SERVICE MANUAL

ASTRO 102BX TRANSCEIVER

CUBIC
COMMUNICATIONS, INC.
SWAN DIVISION

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To the Service Technician:

The ASTRO-102BX is a modern, top-of-the-line, sophisticated Amateur Band Transceiver utilizing state-of-the-art techniques and design. The features that today's Hams want are built into the unit, not marketed as "add on" accessories at extra cost -- dual PTO's, Passband Tuning, Full or Semi Break-in, VOX, CW Sidetone, RIT, Notch Filter, and more.

Whenever performance measured up to required standards, integrated circuit technology was used in the design; where special functions were required, or desired performance could not otherwise be obtained, carefully designed discrete component circuits were included, rather than accept compromised performance.

Every care has been taken to provide long, trouble-free life, but the requirements of extra operating features, high performance design, and compact packaging demanded by today's Hams results in a unit necessitating a high degree of skill on the part of you, the Service Technician. To assist you, this manual has been painstakingly prepared. Take the time to thumb through it; read and study the Theory of Operation section and the Servicing Hints section; it will be time well spent, even for the most highly qualified and experienced personnel.

It is you, the Service Technican, that will ultimately determine how well or how poorly the radio will work, and therefore how satisfied or dissatisfied our customers will be, in the long run. If you have a suggestion or a problem, write or call our service department. We'll listen, and help you all we can.

Sincerely,

CUBIC COMMUNICATIONS, INC.
Swan Division

Sam F. Arn, K6TSD
President

SFA/lb

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Contents

Section One
General Information.............................1.1

Section Two
Specifications..................................2-1

Section Three
Receiver Theory of Operation...............3-1

Section Four
Transmitter Theory of Operation............4-1

Section Five
Oscillator Theory of Operation.............5-1

Section Six
Ancillary Circuits Theory of Operation......6-1

Section Seven
Servicing Hints.................................7-1

Section Eight
Servicing Data..................................8-1

Section Nine
Parts Lists.....................................9-1

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General Information

The Swan ASTRO 102BX is a 160M through 10M SSB amateur transceiver that features dual PTO's and frequency counter for split frequency operation with digital frequency readout. Frequency selection is the only tuning necessary when operating into a matched antenna with a VSWR of 1.7:1 or less.

Standard features include: Full break-in or semi-break-in CW operation; narrow band crystal CW filter; two position CW transmit pulse shaping; VOX; noise blanker; SSB transmit speech processor; true RF and IF gain controls; peak reading wattmeter with FWD, REF, and ALC positions; variable AGC decay; passband tuning; and audio notch filter. The basic single conversion design has been proven in thousands of the popular Swan transceivers. Mechanical and electrical designs are exceptional, for years of trouble-free operation, even in a rugged mobile environment.

The unit is designed for fixed and mobile use with minimum interconnections. For fixed station operation an auxiliary speaker/power supply, model PSU-6, provides 13.6 V DC from 115/230 V AC. A matching antenna tuner, model ST2A, is also available.

The transmitter may be used for RTTY or Slow Scan TV, but operation in these modes requires additional forced air cooling. See the operator's anual for precautions to be observed when operating in these modes. The ASTRO 102BX includes automatic gain control (AGC) in the receive mode, and automatic level control (ALC) in the transmit mode. CW sidetone is also provided.

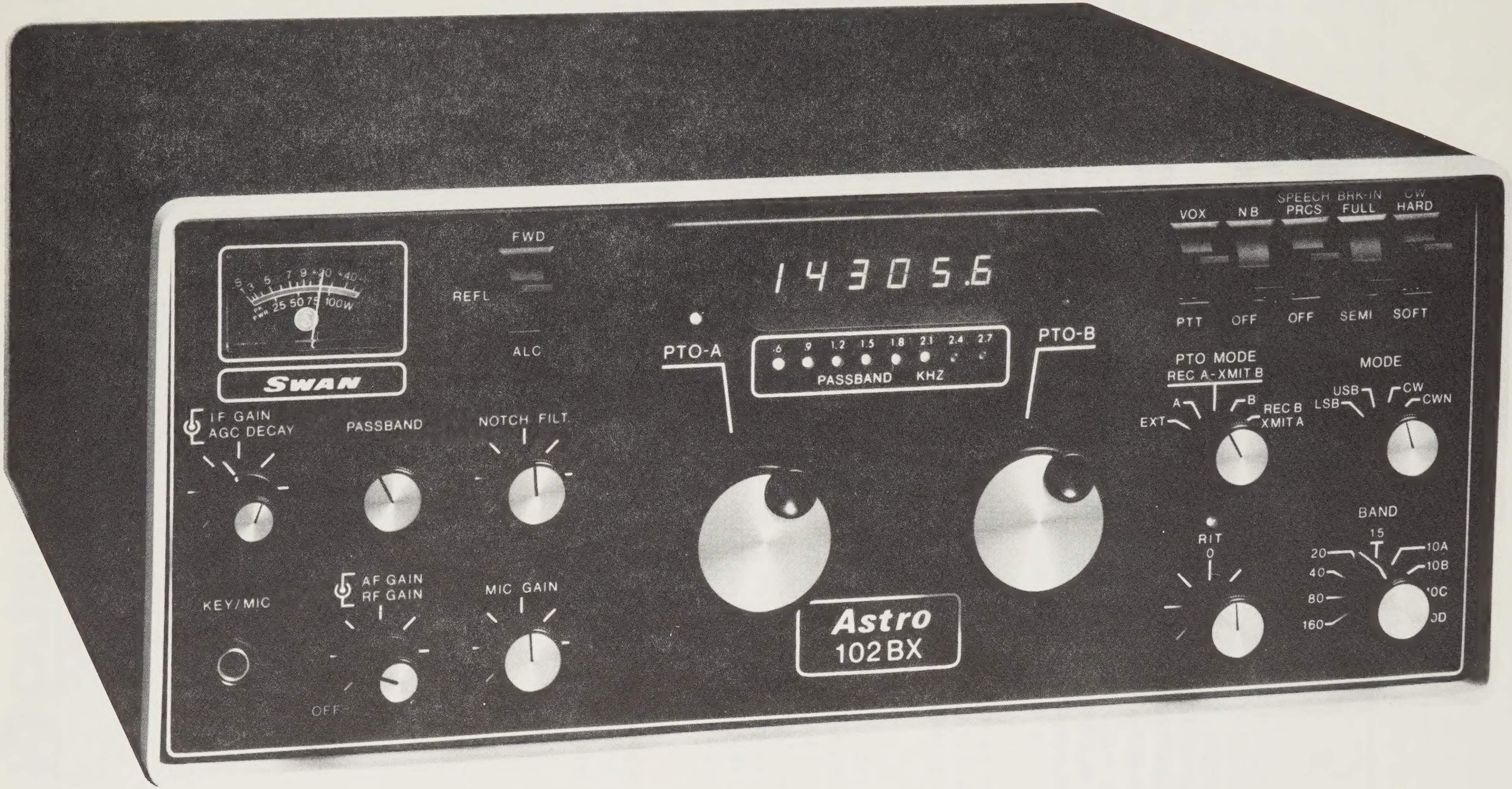

SWAN
FWD
REFL
ALC
14305.6
PTO-A
PTO-B
.6 .9 1.2 1.5 1.8 2.1 2.4 2.7
PASSBAND KHZ
VOX
NB
SPEECH PRCS
BRK-IN FULL
CW HARD
PTT
OFF
OFF
SEMI
SOFT
PTO MODE
REC A-XMIT B
EXT
A
B
REC B XMIT A
MODE
USB
LSB
CW
CWN
IF GAIN
AGC DECAY
PASSBAND
NOTCH FILT.
RIT
0
BAND
15
20
40
80
160
10A
10B
10C
10D
KEY/MIC
AF GAIN
RF GAIN
OFF
MIC GAIN
Astro
102BX

FIGURE 5-1

Specifications

General

Frequency Coverage*

160M	1.8-2.0 MHz
80M	3.5-4.0 MHz
40M	7.0-7.5 MHz
20M	14.0-14.5 MHz
15M	21.0-21.5 MHz
10M	28.0-28.5 MHz 28.5-29.0 MHz 29.0-29.5 MHz 29.5-29.999 MHz

*Approximately 50 to 100 kHz overrange provided on each band.

Frequency Control: Dual PTO with 6-digit frequency counter and RIT control.

PTO Modes of Operation: A only; Receive A-Transmit B; B only; Receive B-Transmit A; External LO.

Modes of Operation: SSB with selectable sideband. CW with automatic 800 Hz offset on transmit and full or semi-breakin.

Power Input Requirements: 12-14 V DC negative ground only. No damage up to 15 VDC. 20 amperes peak current.

Electromagnetic Interference:

Transmit: Five element low pass filter on each band.

Receive: Five element lowpass filter plus three pole bandpass filter on each band.

Dimensions: 6.375" H x 14.25" W x 13.25" D
16.2 cm x 36.2 cm x 33.7 cm
Weight: 23.5 pounds (10.6 kg)

Rear Panel:

Power connector:

Pin 1	13.6 V DC
2	Ground
3	Ground (used for accessories)
4	13.6 V DC
5	Ground
6	13.6 V DC fused to 3 amps (used for accessories)

Antenna connector: UHF (SO-239)

External speaker/phones: Phone jack

External relay: RCA jack

Key:	RCA jack
EXT. Modulation:	RCA jack
Fuse:	3 amp fast blow
Ext. L.O.:	RCA jack

Receiver

Circuit Design:	Single conversion to 9.00165 MHz IF using double balanced mixer. Exceptional immunity to overload and cross modulation.
Receiver Sensitivity:	10 dB S+N/N typical at 0.35 uV
Image Rejection:	Better than 60 dB
Receiver Selectivity:	SSB and CW--2.4 kHz bandwidth, two 8-pole crystal filters with shape factor 1.4; 6 dB to 100 dB CWN--300 Hz bandwidth. IF crystal filter in series with one 8 pole SSB filter
Passband Tuning:	SSB and CW eight pole cutoff continuously variable high pass or low pass. LED readout shows equivalent audio passband. CWN--IF crystal filter continuously tunable over 300 to 3000 Hz with passband control
Dynamic Range:	AGC greater than 100 dB; third order intercept +15 dBm
Audio Output Power:	Greater than 3 watts into 4 ohms.
Meter:	S-units from 1 to 9; 20, 40 and 60 dB over S-9

Transmitter

Circuit Design:	Excellent harmonic and TVI suppression. ALC limits SSB peak power and CW power to 100 watts output.
Power Input:	235 watts PEP input at 13.6 V DC input
Power Output:	100 watts PEP into 50 ohm non-reactive load at 13.6 V DC
Unwanted Sideband Rejection:	>60 dB down at 1000 Hz audio note
Harmonic Output:	>45 dB below peak power level
Carrier Suppression:	>50 dB below peak power level

Spurious Output:	> 55 dB below peak power level
Transmit Control: (for SSB)	PTT and VOX standard
CW Transmit:	Full or semi-breakin with sidetone standard. Automatic 800 Hz offset on transmit frequency.
Microphone Input:	47 kΩ input impedance
Audio Response:	300 to 3000 Hz ± 6 dB
Meter:	In receive mode, reads S units and dB over S-9; in transmit mode, reads peak power selected by meter switch: Forward power 100 watts full scale, reflected power 10 watts full scale, or reads ALC.
Cooling:	Large capacity heat sink fins supplied. For SSTV, RTTY, and semi-continuous transmit, forced air cooling such as a small "muffin" fan on heat sink fins is required.
VSWR Shutdown:	Full power up to VSWR = 1.7:1

Approximate limit ratio as follows:

VSWR	Percent Power
1.7:1	100%
2.0:1	80%
3.0:1	60%
Open or short	25% (equivalent voltage)

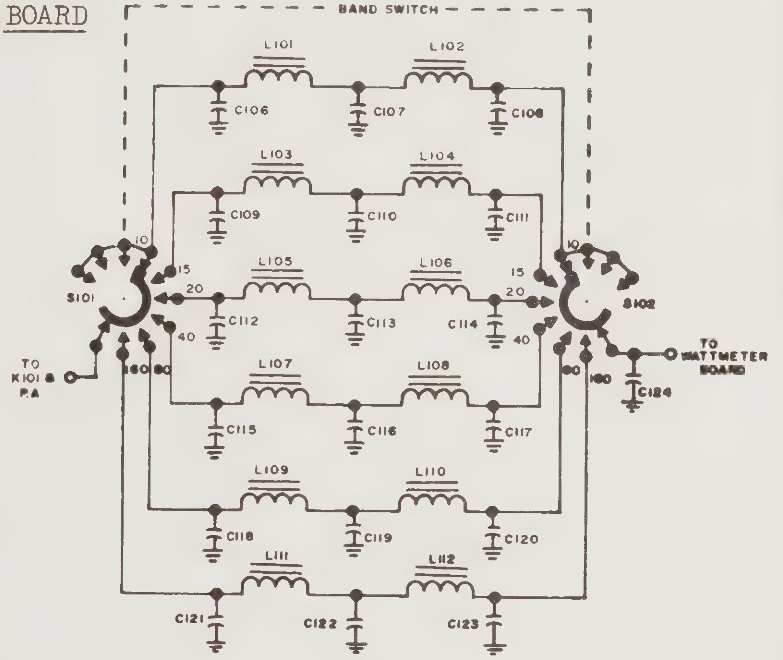
Receiver Theory

Bandswitch and Low Pass Filters.................3-1
Preselector.......................................3-2
First Mixer.......................................3-3
Noise Blanker Gate................................3-4
Filter, Buffer, and Post Filter Amplifier.......3-5
AGC Attenuator....................................3-5
Passband Tuning...................................3-6
CW Narrow Filter..................................3-7
Main I.F. Amplifier...............................3-8
Product Detector/Audio Mute.....................3-8
Notch Filter......................................3-9
Audio Power Amplifier/Receiver Mute.............3-9
Automatic Gain Control.........................3-10

NOTE: Component designations used in the "Theory" sections of this manual are the shortened numbers as they appear on the schematic diagrams. To determine complete component designation, the board/location number must be added after the letter or letters. See the "Note" in the first portion of the "Parts" section of this manual.

Schematic diagrams contained in the "Theory" sections of this manual are simplified to help in the understanding of the operation of the circuit; they may not always, therefore, be sufficient for troubleshooting purposes. When servicing the equipment, it is urged that the technician use only the schematic diagrams included in the "Servicing Data" portion of this manual.

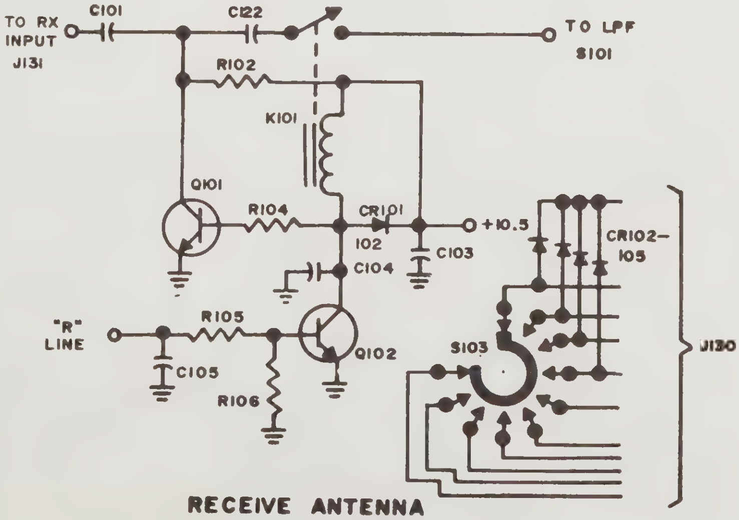
BANDSWITCH AND LOW PASS FILTER BOARD



The received signal from the antenna connector is passed through one of six low pass filters, which are also used in the transmit condition. Each is composed of two inductors and three capacitors, yielding a five pole filter for each band (160 through 10 meters). Note that the same filter is used for all segments of the 10 meter band.

The filters are selected by S101 and S102, sections of the BAND switch. From S101, the signal is passed to relay K101, through C122, and through C101. From there it is routed to the input of the preselector board.

Receiver antenna switching is accomplished by K101, which is activated by Q102. When the receive ("R") line is high, as it is in the receive condition Q102 is biased to saturation, energizing K101 and connecting C122 to the output of the selected low pass filter. Since the collector of Q102 is low when it is thus saturated, Q101 is in a non-conducting condition and has no effect. However, when the "R" line is low, as



RECEIVE ANTENNA SWITCHING

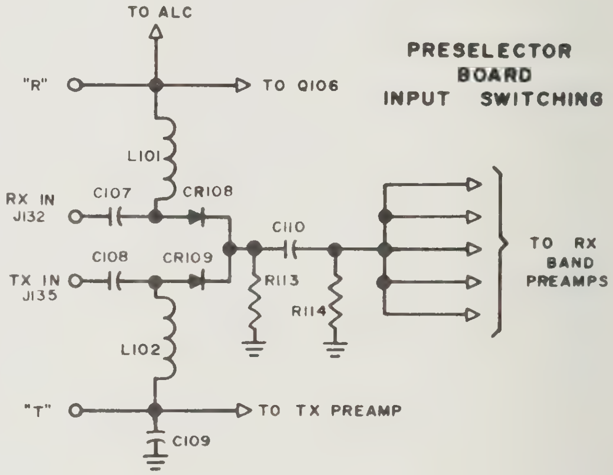
as it is in the transmit condition, Q102 is cut off, and its collector voltage is high. This biases Q101 on, providing a path to ground for any transmit signal coupled across the open contacts of K101 due to stray capacity. In this manner, the input of the receiver is protected during transmission periods.

S103, a part of the BAND switch, is used to provide a voltage to each of ten control lines, depending upon its position. There are separate control lines for the 160, 80, 40, 20, 15, 10A, 10B, 10C and 10D positions, and an additional line, which, through CR102 through CR105, is high for any of the 10 meter positions of the BAND switch. These control lines are used in other sections of the transceiver.

PRESELECTOR

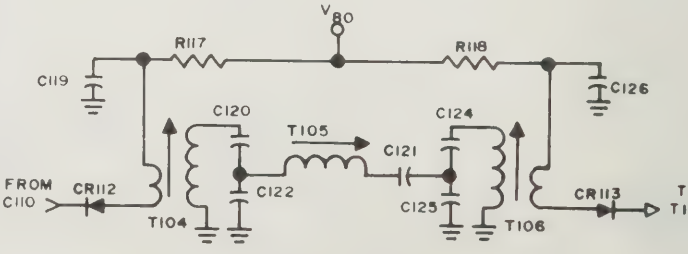
In the receive condition, the transmit ("T") line is low. Thus, CR109 in the input circuitry of the preselector is reverse biased, and the TX IN jack is effectively open circuited. L102 is an RF choke to prevent signal from being bypassed to ground through C109 on the "T" line.

On the other hand, the "R" line is high in the receive condition, and is connected through RF choke L101 to CR108, which it biases to an "on" condition. R113 provides a ground return and the voltage drop across it reverse biases the "off" diode. C110 is a DC blocking capacitor, and passes the signal on to the input of the preselector band filters. R114 acts as a DC return for the switching currents for the filter selection circuits, and provides reverse bias to the "off" diodes.


PRESELECTOR BOARD INPUT SWITCHING

There are six band filters in the preselector (only one is used for all four segments of the 10 meter band). These filters are selected by the BAND switch, utilizing the voltages on the control lines from S103, as mentioned earlier. For discussion purposes, only the 80 meter filter will be considered.

The voltage from the 80 meter control line is coupled through isolating resistor R117, the primary of T104, and CR112 to the output of the input switching circuit previously discussed, where it is provided with a ground return path through R114. C119 provides a ground for the signal at one end of the primary of T104. The signal is then passed through diode switch CR112 to T104, while the voltage drop across R114 reverse biases the corresponding diode switches in the other filters.

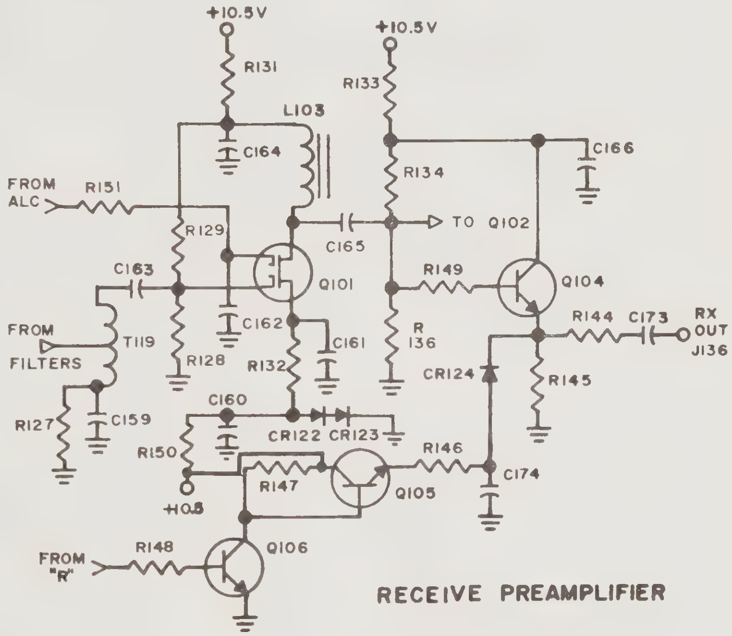

80 M BAND FILTER

The secondary of T104 is resonated by the C120-C122 series combination, and they act as an impedance transformation voltage divider. T105 and C121 form a series resonant circuit, and its output is coupled to C124-C125, which again transforms the impedance of the signal path, and also resonates with output transformer T106. The combination of the two parallel resonant circuits and one series resonant circuit forms a bandpass filter of excellent characteristics. Output switching is provided by R118, in conjunction with CR113, in a manner identical to the input switching scheme.

The output of the preselector filters is coupled to the preamplifier stage, Q101, via autotransformer T119 and coupling capacitor C163. R127 provides a DC path to ground for the switching diode, and provides reverse bias for the "off" diodes in the other filters, while C159 effectively grounds the bottom of T119 for the signal.

Bias for Q101 is provided both by the source circuitry and the R128-R129 combination. So far as gate two is concerned, the bias is determined by the source network as well as whatever is applied to it via the ALC line. CR122 and CR123, biased in the forward direction by the voltage applied by R150, provide a fixed drop of approximately 1.4 volts, while additional bias is supplied by the drop across source resistor R132. C160 and C161 are both bypasses.

Signal from the drain of Q101 is taken via C165, after being developed across load L103. R131 and C164 form a decoupling network for the supply line.



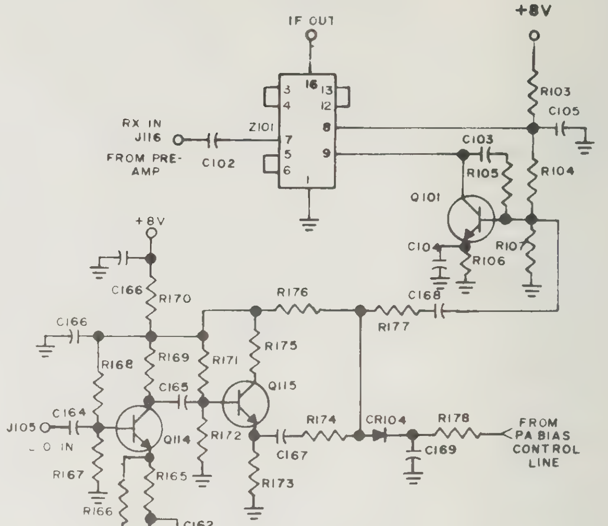
RECEIVE PREAMPLIFIER

In the receive condition, the "R" line is high, causing Q106 to saturate, thus lowering the voltage on the base of Q105, causing it to cease conduction. Diode switch CR124 is then reverse biased by the voltage drop across R145, allowing Q104 to operate in a normal manner. Bias for Q104 is derived by R134 and R136, while R149 serves to prevent instability. Output of the preselector is taken from the emitter of Q104 via R144 and C173 and coupled to the receiver input of the exciter board.

EXCITER BOARD--FIRST MIXER

After passing through the preselector and being amplified by the preamplifier, the received signal is connected to mixer Z101 via C102, a blocking capacitor. Local oscillator signal, brought to the exciter board through J105, is coupled via blocking capacitor C164 to the base of amplifier Q114. Bias for Q114 is developed in a conventional manner via the R167-R168 voltage divider. The emitter resistor of Q114 is divided into two parts, R164 and R165. While the former is bypassed in the usual manner, the latter is not, and a signal suitable for driving the display frequency counter is taken from across it, via R166 and C163.

Amplified L.O. output is taken from the collector of Q114 and coupled to the base of Q115. Again, bias is conventional, being developed by R171 and R172. The output of Q115 is taken across its emitter resistor, R173, via C167 and R174, and coupled through R177 and C168 to the base of mixer driver Q101. The P.A. Bias control line is high in the receive condition, thus reverse biasing CR104, preventing the L.O. signal from being shunted to ground by C169, as it would be in the transmit condition.

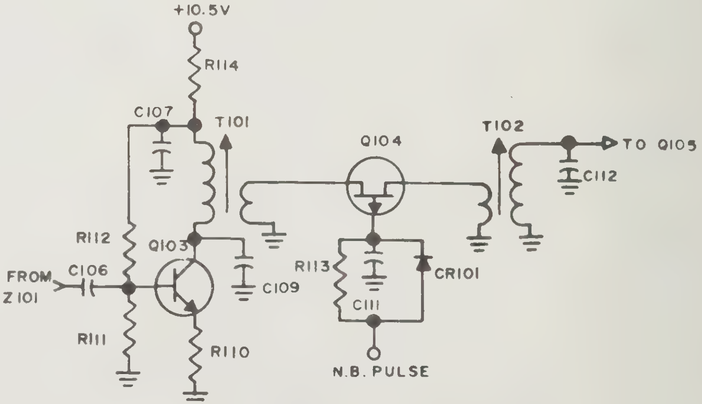


L.O. AMPLIFIERS & 1ST MIXER

Bias for Q101 is developed by R104 and R107, and C104-R106 form the usual emitter circuit. C103 and R105 form a feedback network, which, in conjunction with R177 and R174 set the gain of Q101. For a more complete discussion of feedback amplifiers and their operation, refer to the "Servicing Hints" section of this manual.

NOISE BLANKER GATE

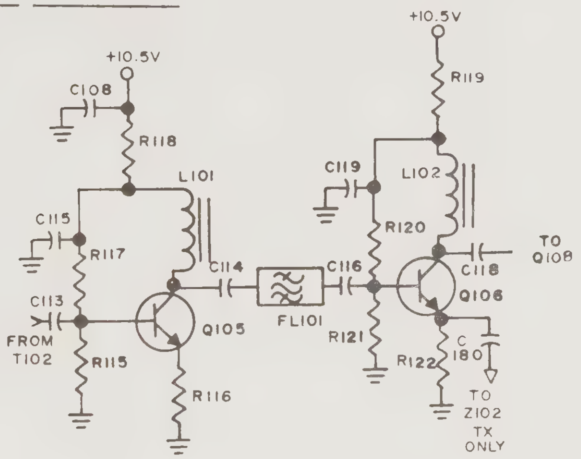
The output of the first mixer, Z101, now at the I.F. frequency, is coupled to the first I.F. amplifier, Q103, via blocking capacitor C106. R111 and R112 form the bias network, while emitter resistor R110 is unbypassed, providing some degenerative feedback. Output is developed across a parallel-tuned resonant circuit composed of the primary of T101 and C109. R114 and C107 form a decoupling network, while the latter also places the upper end of the primary of T101 at AC ground, so that C109 is effectively in parallel with T101.



FIRST I.F. AMPLIFIER & NOISE BLANKING GATE

The secondary of T101 presents a signal at low impedance, which is coupled through noise blanker gate Q104 to the primary of T102, where it is stepped back up. C112 resonates the secondary of T102. When the noise blanker circuit detects a noise pulse of sufficient amplitude, the gate of Q-104 is fed a positive pulse through CR101, effectively cutting it off and opening its source-to-drain path, thus opening the signal path. Otherwise, a ground is applied to the gate via R113, keeping Q104 in an "on" condition. C111 and R113 serve to lengthen the pulse at the gate of Q104 to ensure positive blanking.

FILTER BUFFER, FILTER, AND POST FILTER AMPLIFIER



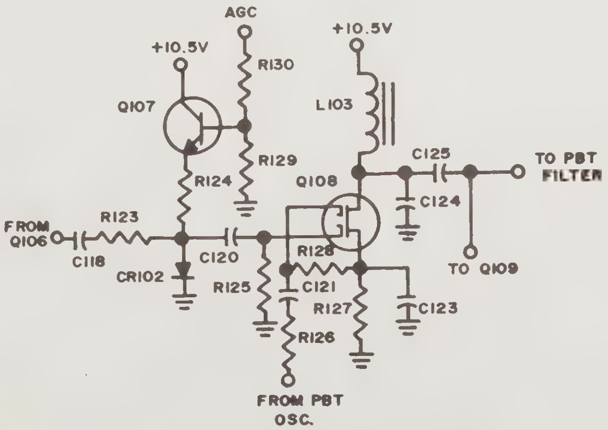
FILTER BUFFER, FILTER, AND
POST FILTER AMPLIFIER

From T102, the signal is coupled to filter buffer amplifier Q105 via blocking capacitor C113. Bias is developed by R115 and R117, while emitter resistor R116 is unbypassed, providing degeneration. Output is taken across collector load L101 and coupled through blocking capacitor C114 to the first I.F. filter, FL101. R118, C108, and C115 are a decoupling network.

From filter FL101, the signal is coupled through C-116 to the base of post filter amplifier Q106, which is biased by R120 and R121. In the transmit condition, Q106 acts as an emitter follower (discussed in detail in the Transmit section of this manual), while in the receive condition, the output is taken from the collector after being developed across load L102. C118 is a blocking capacitor. Note that the use of inductive loads and unbypassed emitters in the stages preceeding the AGC attenuator ensure good strong signal handling capability.

AGC ATTENUATOR

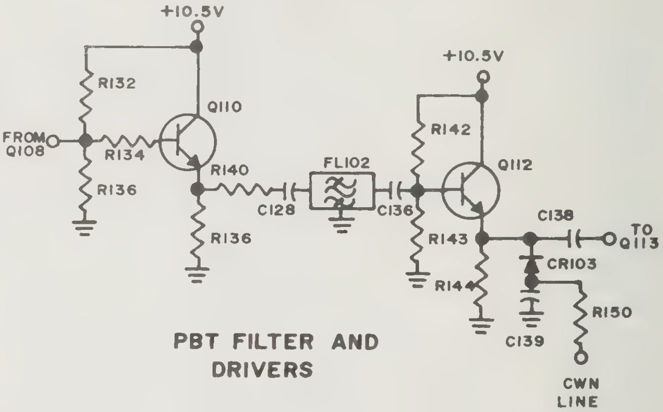
From Q106, the received I.F. signal is coupled through C118 and R123 to AGC attenuator PIN diode CR102, and thence via C120 to gate one of pass band tuning first mixer, Q108. AGC voltage, when required, is applied to the base of AGC driver Q107 via voltage divider R129-R130. Current is then supplied to the PIN diode via R124, causing it to conduct to a degree determined by the AGC voltage; it acts, in conjunction with R123, as a variable voltage divider, similar in effect to a conventional volume control, and thus provides the AGC action required. Up to about 50 dB of attenuation is available at this point.



I.F. AGC ATTENUATOR AND
PBT 1st MIXER

PASSBAND TUNING

Q108, a dual-gate MOSFET, acts as a subtractive mixer, with the first I.F. signal applied to gate one, and the passband tuning oscillator signal applied to gate two. Bias for gate one is developed across R127, while connecting gate two to the source via R128 makes it operate without bias. Output of Q108, now at the pass band tuning I.F. of 13.8 MHz, is taken from its drain. L103 and C124 form a parallel resonant drain load. C125 is a coupling capacitor, routing the signal to the SSB filter via Q110 and to the CW narrow filter via Q109.

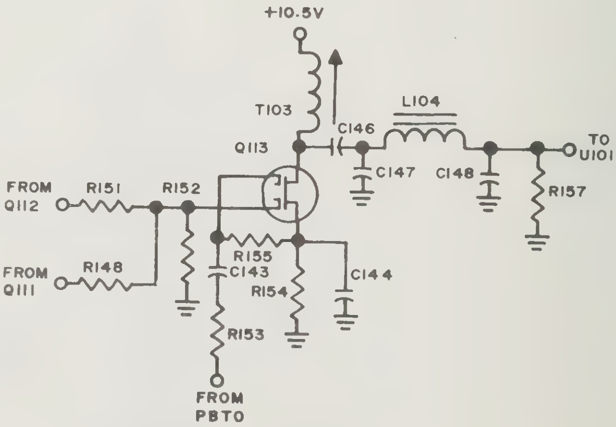


PBT FILTER AND DRIVERS

Q110, conventionally biased, acts as an emitter follower to drive the SSB filter, FL102. After passing through the filter, the signal is coupled to Q112 via C136. Q112 is also an emitter follower, with its output developed across R144 and bias developed by the R142-R143 combination. C138 is a blocking capacitor.

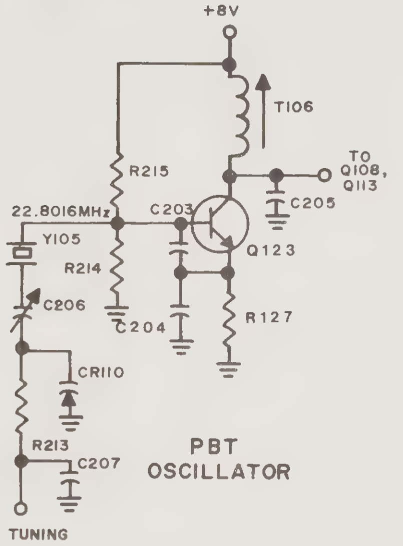
When operating in the CW narrow mode, voltage is applied to the CWN line through R150 to diode switch CR103. This voltage reverse biases the base-emitter junction of Q112, causing it to cease to conduct, and thus breaking the signal path. Also, C139 shorts the signal path through CR103.

Signal from the CWN filter amplifier Q111 (covered later) and from the SSB filter are coupled to gate one of the second PBT mixer Q113. R148 and R151 are isolation resistors, and R152 is the gate return resistance for Q113, which is also a subtractive mixer. Oscillator energy is coupled to gate two of Q113 in the same manner as it was to the first PBT mixer, and the bias arrangement is also the same. Output is taken via C146, now back at the first I.F. frequency of 9 MHz. C147 serves the dual function of resonating with T103 and serving as the first element in a low pass filter composed of C147, L104 and C148, which filters out unwanted mixer products. R157 terminates the filter in its required load.



2nd PBT MIXER

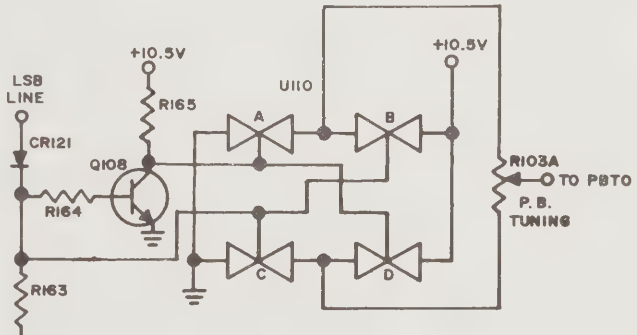
PBT oscillator Q123 is a conventional Colpitts crystal controlled oscillator, with C203 and C204 providing the feedback capacitive divider. R214 and R215 form the bias network, while R127 is the emitter resistance. T106 and C205 form a parallel resonant collector load. The frequency of the oscillator is adjusted by the application of a voltage from the PASSBAND tuning control on the front panel through isolation resistor R213, to varicap CR110. The range of adjustment is set by trimmer C206, while C207 is a bypass.



PBT
OSCILLATOR

Because the same frequency oscillator signal is applied to both PBT mixers, any signal going into the first PBT mixer will leave the second on precisely the same frequency. This is why the PASSBAND tuning control only affects the relative position of the passband and has no effect on the frequency of the signal tuned in.

The direction of tuning of the passband oscillator must be one way for USB/CW operation, and the opposite direction for LSB operation for the effect at audio to remain the same. When in the LSB mode, a positive voltage is applied through CR121 on the audio board to the base of Q108, through limiting resistor R164. Simultaneously, it is also used to turn sections B and C of switch U110 on. Section B of U110 applies +10.5 volts to the upper end of the PASSBAND tuning control R103A, while section C grounds the other end. Since Q108 is saturated, its collector is low, and this voltage, applied to sections A and D of U110, turns them off. When in the USB or CW modes, however, the LSB line is low, and thus sections A and D of U110 are on, while sections B and C are off. This action reverses the voltage applied to the tuning control.

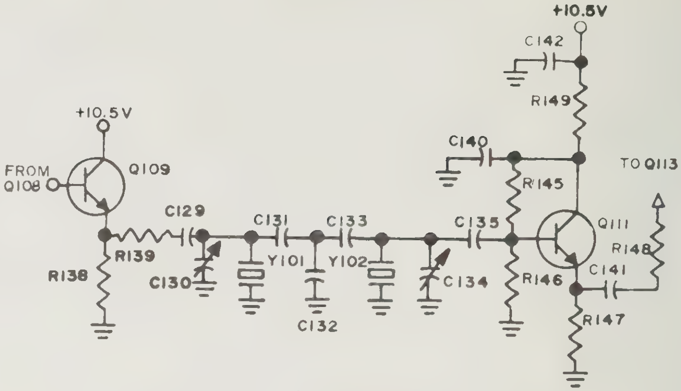


PBT SWITCHING

Operation of the PASSBAND tuning indicator will be covered in the "Ancillary Circuits" section of this manual.

CWN FILTER

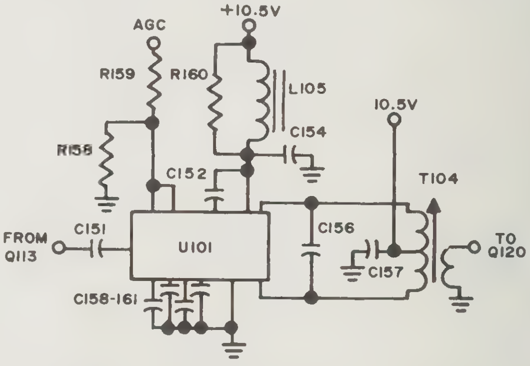
Signal at the second I.F. of 13.8 MHz, from the first passband tuning mixer (Q108), is also coupled to the base of Q109. Bias is received from the same network that supplies Q110, the filter driver. Q109 acts as an emitter follower driver for the CW narrow filter, and the required filter termination impedance is provided by R138 and R139, while C129 is a blocking capacitor. The CW narrow filter is composed of Y101 and Y102, with C131 and C133 acting as coupling capacitors and C130 and C134 providing adjustments to properly align the filter response. C135 acts as an output coupling capacitor, and carries the signal to the base of Q111. Power for Q111 is taken from the CWN line, and bias is provided by R145 and R146. Output is taken across emitter resistor R147 and coupled to the second passband tuning mixer Q113 through C141 and R148.



NARROW CW FILTER

MAIN I.F. AMPLIFIER

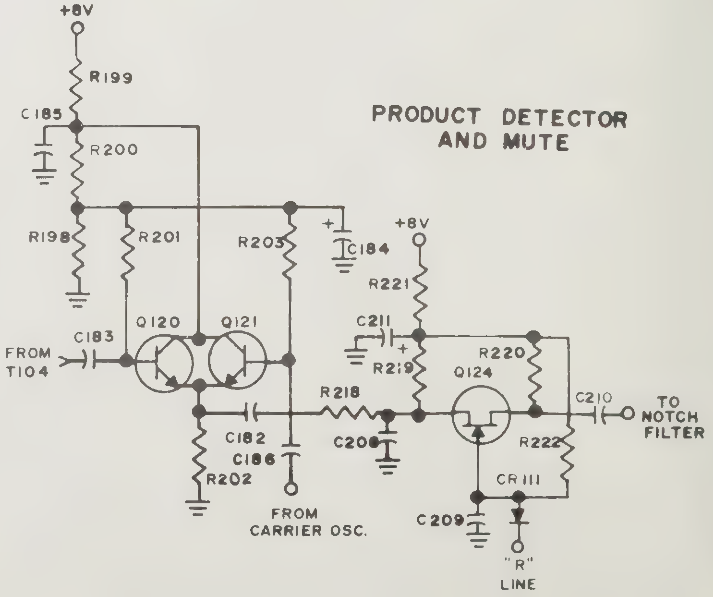
The signal from Q113, now again at the 9 MHz I.F., is coupled through C151 to the main I.F. amplifier, U101. This integrated circuit is composed of two amplifier sections, with the output of the first section developed across the R160-L105 load combination. C154 resonates with L105, while R160 broadens the response by "swamping" the tuned circuit. C152 couples the output thus developed from the first amplifier section of U101 to the second section of the same IC. Output of the second section is coupled to push-pull T104, which is resonated by C156. C157 is a bypass. The output of T104 is coupled to the product detector, Q120. AGC control is applied through voltage divider R158—R159. Overall gain of U101 is up to 70 dB.



MAIN I.F. AMPLIFIER

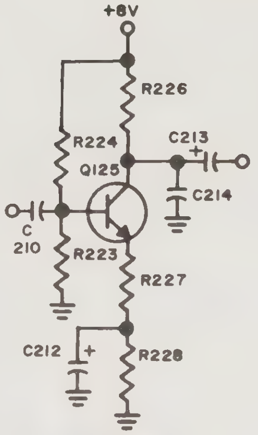
PRODUCT DETECTOR/AUDIO MUTE

Carrier oscillator energy is coupled via C186 to the base of Q121, which forms an emitter-coupled pair with Q120. Bias is developed by R200 and R198, acting as a voltage divider, and R201 and R203 serve as isolating resistors, while C184 is a bypass. Q120 and Q121 act as a mixer, combining the I.F. input from T104 and the carrier oscillator signals to recover the audio frequency of the incoming signal. Output is coupled via C182 to an RC network composed of R218 and C208,



PRODUCT DETECTOR AND MUTE

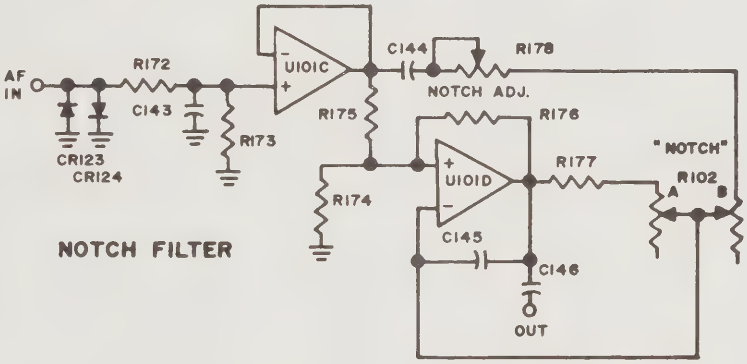
which filters out any undesired mixer products. Q124, in series with the signal path, acts as a switch to cut off the receiver output AF when in the transmit condition. In that instance, the "R" line would be low, grounding the gate of Q124 and causing it to act as an open circuit. R219 and R220 provide bias voltage for the source and gate. Output of the AF mute gate is then coupled via C210 to the base of AF amplifier Q125, which provides additional gain. Bias is conventional, while the emitter resistance is only partially bypassed to set the gain (see the "Servicing Hints" section of this manual for a discussion of feedback amplifiers such as this). Output is taken via C213, while C214 provides a degree of high frequency rolloff for more pleasing audio response.



A.F. PREAMP

NOTCH FILTER

Audio from Q125 is applied to CR123 and CR124, which act as a limiter, effectively clipping off any peaks exceeding approximately ± 0.7 volt. The R172-C143 combination filters the output of the clipper and shapes the audio response. U101C is connected as a voltage follower, providing drive for active filter U101D that is the same amplitude and phase as its input, but of a lower driving impedance.

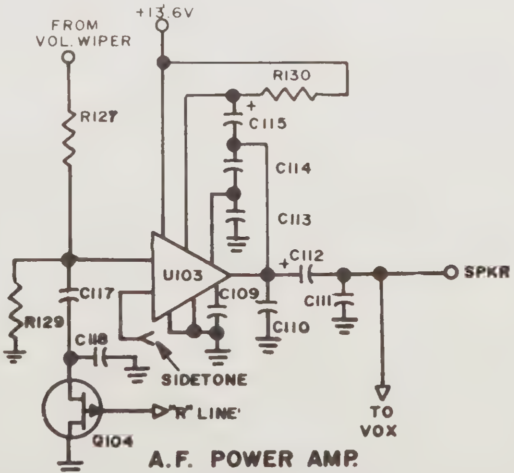


NOTCH FILTER

C144 and R178, with R120A on the front panel, form the series leg of a Wein bridge, while C145, R177 and R102B (also on the front panel) form the shunt leg of the same Wein bridge. This bridge is inserted in the inverting feedback loop to form a notch when combined with the normal audio frequency applied to the non-inverting input of amplifier U101D. R176 and R174 form a positive feedback path to raise the"Q" of the circuit.

AUDIO POWER AMPLIFIER/RECEIVER MUTE

Audio from the output of the notch filter is applied to the high side of the front panel volume control, and its wiper is returned, through R127, to the input of audio power amplifier U103. R129 acts in conjunction with R127 as a voltage divider. Receiver mute switch Q104 is connected between this input and ground, through isolating capacitor C117. When in the receive condition, a voltage on the "R" line holds Q104 in the "open" condition, allowing the signal to be amplified by U103; when in the transmit condition, the "R" line is grounded, turning Q104 "on" and shunting the signal to gr-



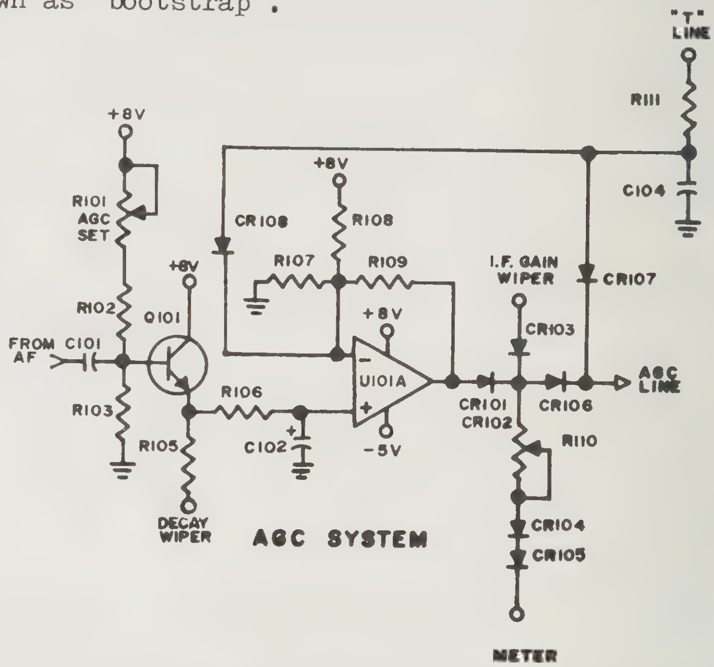
A.F. POWER AMP.
AND RECEIVER MUTE

ound. C118 and R127 shape the audio response.

When in the transmit condition, and when sending CW, the sidetone audio signal is applied to the other input of U103; thus, the operator is able to monitor his own sending.

C113 and C114 form an audio shaping network for more pleasing response, and C112 couples the output to the speaker. C110 and C111 are RF bypasses. C109 bypasses certain internal circuitry of U103, while R130 and C115 provide a type of feedback path known as "bootstrap".

AUTOMATIC GAIN CONTROL



AGC SYSTEM

The audio frequency signal is coupled to the base of Q101 via C101. Bias is set by R101, R102 and R103, with R101 serving as the AGC threshold adjustment. When the audio peaks are of sufficient amplitude, Q101 conducts, which causes a current to flow through R105 and the DECAY control, a variable resistance to ground. The voltage drop across this combination charges capacitor C102, with series resistor R106 controlling the charging time, and thereby the attack time. When the AF signal disappears, C102 discharges through R106 and the R105-DECAY control combination, thus the setting of the control varies the decay time of the AGC.

The voltage across C102 is applied to the non-inverting input of amplifier U101A. Provided the output of U101A is greater than any reverse bias to CR101 supplied through CR103 by the I.F. GAIN control, the output of U101A is applied to the AGC line through CR106, and to the S meter through calibration control R110 and voltage dropping diodes CR104 and CR105. R107, R108 and R109 establish a reference level and the gain of U101A.

When in the transmit condition, the "T" line goes high, and applies a voltage through CR108 to the inverting input of U101A, causing its output to go low and remain there so long as the "T" line remains high, thus disabling the amplifier. This ensures that the S meter is disabled in transmit. This same voltage is applied, through CR107, to the AGC line to force maximum attenuation of the AGC PIN diode in the IF amplifier chain, and to cut off the I.F. integrated circuit amplifier, U101, on the exciter board. This ensures that no signal is present in the AGC loop during transmit and full gain is available when switched back to receive.

Transmitter Theory

Microphone Preamplifier.................4-1
Speech Processor........................4-1
Modulation Amplifier....................4-1
Balanced Modulator......................4-2
First I.F. Amplifier....................4-2
Noise Blanker Gate......................4-2
Filter Buffer, Filter...................4-3
Transmit Mixer..........................4-3

Preselector

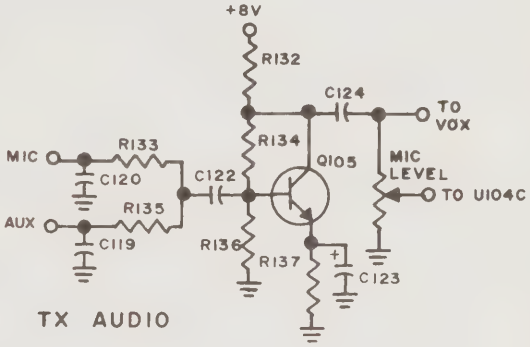
Input switching.......................4-4
Bandpass Filters......................4-4
Amplifiers............................4-5

Power Amplifier

Predriver.............................4-6
Driver Amplifiers.....................4-6
Output Amplifiers.....................4-7
P.A. Bias Control.....................4-7
Low Pass Filters........................4-7

MICROPHONE PREAMPLIFIER

The audio signal from the microphone or external input jack is coupled through R133 or R135, respectively, to blocking capacitor C122 and thence to the base of microphone preamplifier Q103. C119 and C120 are RF bypasses, while bias is developed from the collector by voltage divider R134-R136. This provides feedback to keep the gain of the stage fixed. The emitter circuit is standard, composed of resistor R137 and bypass capacitor C123. Output is coupled via C124 to the VOX circuit and to the internal microphone level control.

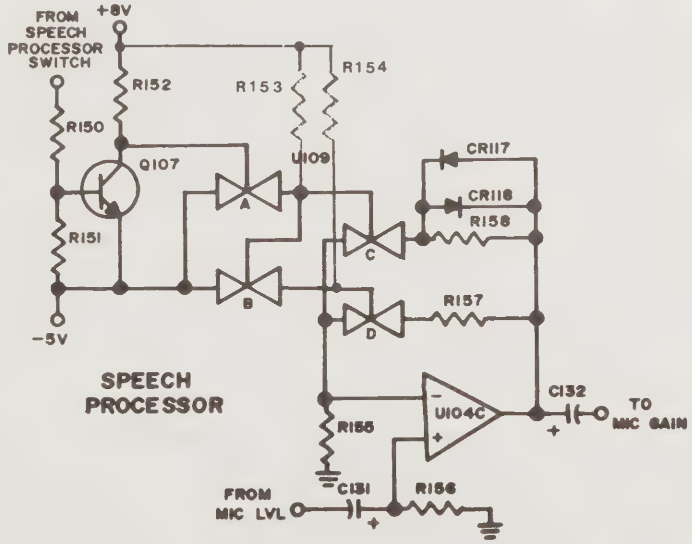


TX AUDIO

SPEECH PROCESSOR

When the operator desires to use the built-in speech processor, a voltage is applied, through voltage divider R150-R151, that biases Q107 into conduction thus lowering its collector voltage and turning section A of switch U109 off. This in turn activates sections B and C of that switching IC. The former turns section D off, opening the feedback path for U104C composed of R157, and the latter completes the alternative feedback path composed of R158, CR117 and CR118.

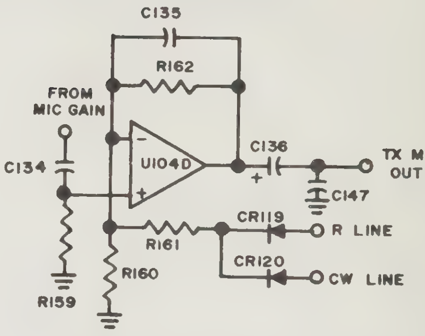
U104C, in this case, operates as a compressor. When signal peaks at the output of the amplifier exceed the barrier potential of CR117 and CR118 (± 0.7 volt) these peaks are fed back to the inverting input of that amplifier, thus reducing its gain. When speech processor operation is not desired, R157 is utilized as the feedback path, converting U104C into an ordinary fixed-gain amplifier stage.



SPEECH PROCESSOR

MODULATION AMPLIFIER

U104D is a fixed-gain amplifier, immediately following the front panel MIC GAIN control. Feedback is provided by the R162-C135 parallel combination, with C135 providing additional feedback at the higher audio frequencies thus reducing gain of the stage in that range, and providing a more pleasant normal sound to the transmitted voice signal. The output is taken via C136, with C147 providing bypassing for RF energy. When operating in the CW mode, or in the receive condition, the amplifier is disabled by voltages applied through CR120 or CR119, respectively, and through R161, to the inverting input of the amplifier, causing its output to go to its lowest limit and remain there so long as these voltages are present, preventing it from amplifying the transmit audio signal.

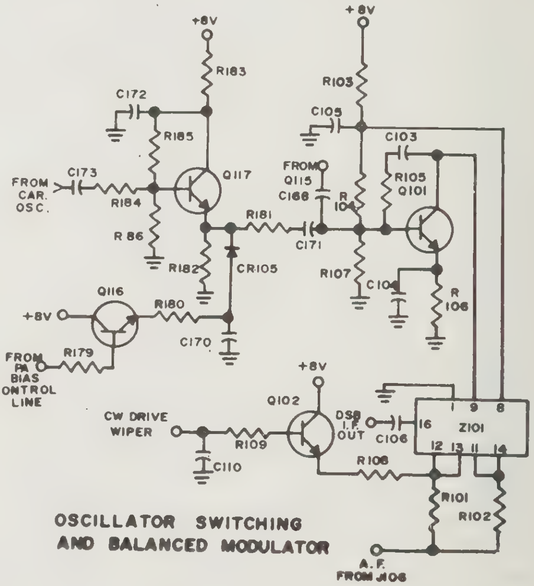


MODULATION AMPLIFIER

BALANCED MODULATOR

Z101, on the exciter board, used as the first mixer in the receive condition, is used as the balanced modulator in the transmit condition. Audio from the modulation amplifier is applied to Z101 through R101 and R102. Meanwhile, the P.A. bias line, which goes low for transmit, cuts off Q116, removing the voltage applied through R180 and CR105 from the emitter of Q117 which allows it to operate as a normal amplifier. This voltage, present in the receive condition, reverse biases the base-emitter junction of Q117, disabling it when it is not needed.

Carrier oscillator energy is coupled through C173 and R184 to the base of Q117, where it is amplified. Output is taken from the emitter, through R181 and C171, and applied to the base of Q101. (In the receive condition, local oscillator energy is coupled through Q115 to the base of Q101, but this path is interrupted in the transmit condition). Q101 is a feedback amplifier, with its gain determined by the ratio of feedback resistor R105 to the input series resistor, R181. Bias and emitter circuits are conventional, and output is taken from the collector of Q101 through a winding internal to Z101. The DSB suppressed-carrier output is taken through C106. (For internal circuitry of Z101, see the Servicing Hints section of this manual.)

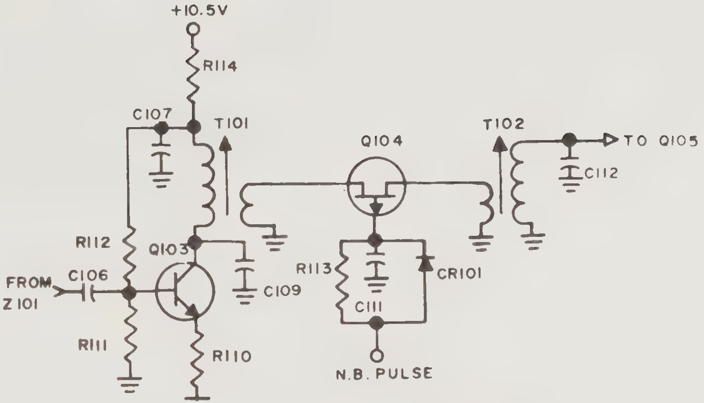


OSCILLATOR SWITCHING AND BALANCED MODULATOR

FIRST I.F./NOISE BLANKER GATE

The DSBSC output of Z101 is coupled to the first I.F. amplifier, Q103, via C106. R111 and R112 form the bias network, while emitter resistor R110 is unbypassed, providing some degenerative feedback. Output is developed across a parallel-tuned resonant circuit composed of the primary of T101 and C109. R114 and C107 form a decoupling network, while the latter also places the upper end of the primary of T101 at AC ground.

The secondary of T101 presents a signal at low impedance which is coupled through noise blanker gate Q104 to the primary of T102, where it is stepped back up. C112 resonates the secondary of T102. In the transmit condition, the noise blanker gate is held in the on condition at all times.

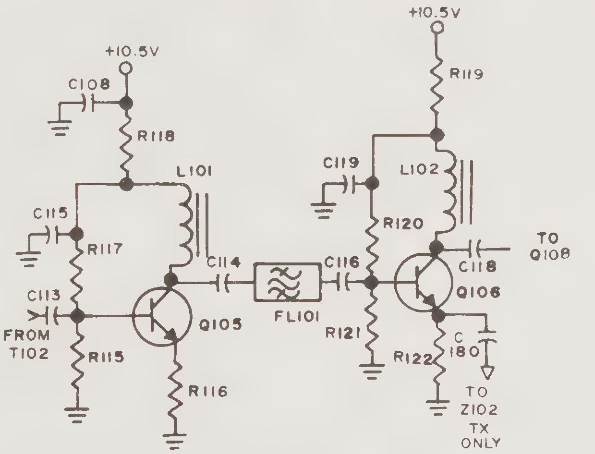


FIRST I.F. AMPLIFIER & NOISE BLANKING GATE

FILTER BUFFER AND FILTER

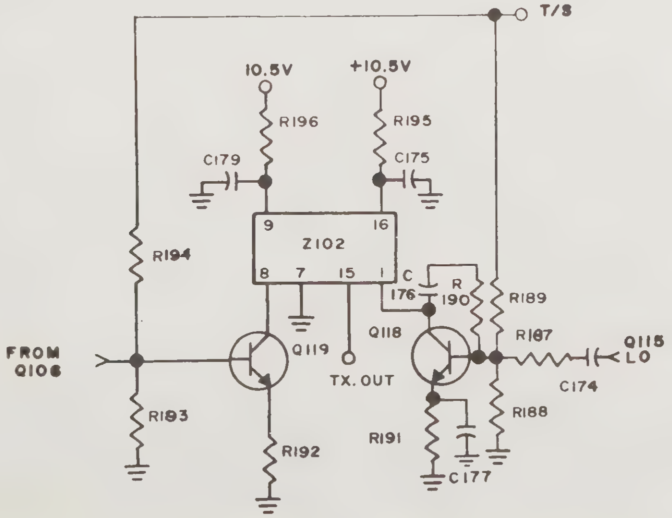
From T102, the signal is coupled to the filter buffer amplifier, Q105, via blocking capacitor C113. Bias is developed by R115 and R117, while emitter resistor R116 is unbypassed. Output is taken across collector load L101 and coupled through blocking capacitor C114 to the first IF filter, FL101. R118, C108 and C115 are a decoupling network.

From filter FL101, the signal is coupled through C116 to the base of post filter amplifier Q106, which is biased by voltage divider R120 and R-121. In the transmit condition Q106 acts as an emitter follower, with its output coupled via C180 and R197 to the transmit mixer driver, Q119.



TRANSMIT MIXER

As mentioned above, the signal from the post filter amplifier, Q106, which is now a single sideband, suppressed carrier signal at the IF frequency of 9 MHz, is coupled to transmit mixer driver Q119. This stage has conventional bias, an unbypassed emitter, and its output taken from the collector through a winding of Z102. This winding is decoupled by R196 and C179. Bias supply comes from the T/S line from the control circuitry,



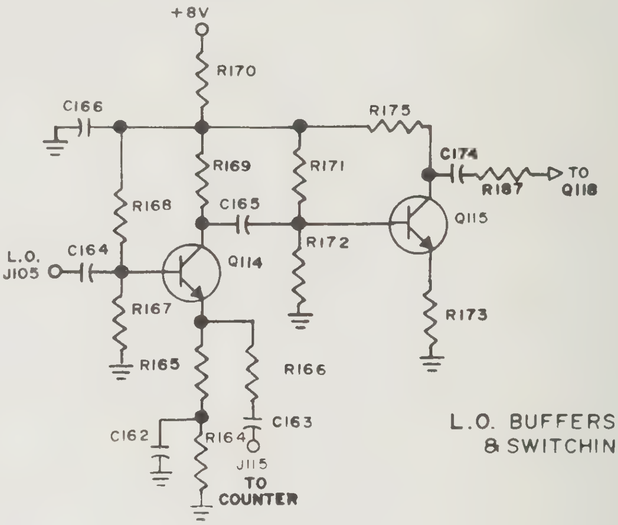
TX MIXER, DRIVERS

which will be covered later.

Local oscillator energy is coupled to the base of the other driver, Q-118. The gain of this stage is determined by the ratio of feedback resistor R190 to input resistor R187; C174 and C176 are merely DC blocking capacitors. The emitter of the stage is bypassed conventionally. Output is taken from the collector through a winding of the transmit mixer, Z102, a doubly balanced mixer assembly identical to that used for A101.

L.O. drive is connected to Z102 through an amplifier/buffer network contained on the exciter board also. Signal from the local oscillator board is applied to buffer Q114 via C164. Output is taken from the R165 portion of the emitter resistance, and coupled through R166 and C163 to the counter board, to provide the signal for the front panel digital dial, while the remainder of the emitter resistance, R164, is bypassed by C162.

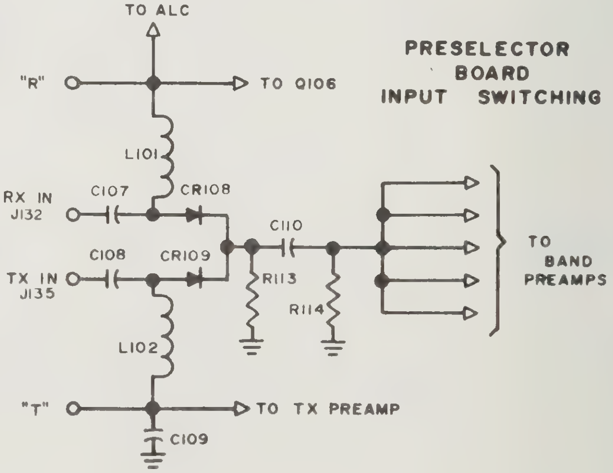
Signal is also taken from the collector of Q114, coupled through C165, and applied to the base of Q115. The emitter resistor of this stage is unbypassed, creating some degree of degenerative feedback, and the output is taken via C174 and R187 to the mixer driver, Q118.



L.O. BUFFERS & SWITCHING

PRESELECTOR--INPUT SWITCHING

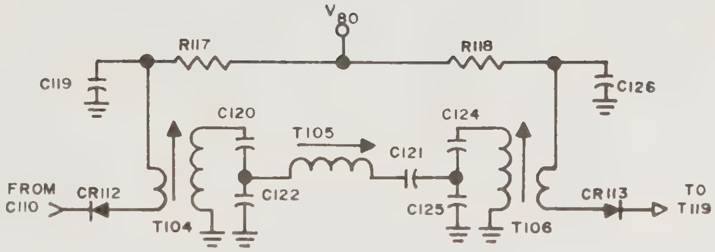
Output from transmit mixer Z102 which is now at the desired operating frequency, is coupled to the same preselector board used in the receive portion of the transceiver. It is coupled via C108 and C110 to the input of the band preamplifier filters when the "T" line is high, biasing switch CR109 on through L102 and R113. Since the "R" line is low under transmit conditions, the voltage dropped across R113 reverse biases receive switch CR108, disconnecting the receiver input from the preselector.



PRESELECTOR BOARD INPUT SWITCHING

PRESELECTOR--BANDPASS FILTERS

Since operation of all of the individual band filters is identical, only one, the 80 meter version, is discussed here. These filters are selected by the BAND switch on the front panel of the transceiver, utilizing the control lines from S103 on the Bandswitch/Low Pass Filter board.

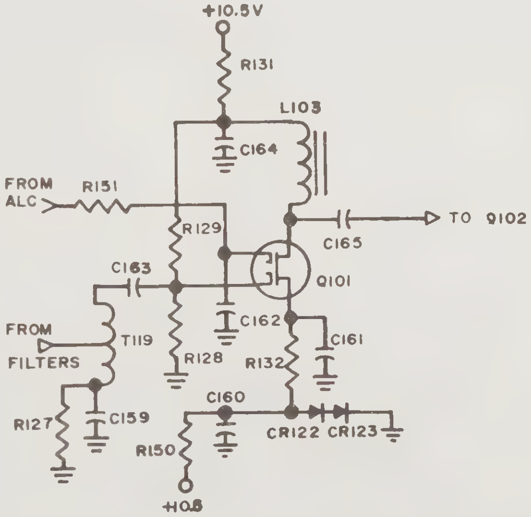


80 M BAND FILTER

The voltage from the 80 meter control line is coupled through isolating resistor R117, the primary of T104, and CR112 to the output of the switching circuitry discussed in the previous section, where it is provided with a ground return path through R114. C119 provides a ground for the signal at one end of the primary of T104. The signal is then passed through diode switch CR112 to T104, while the voltage drop across R114 reverse biases the corresponding diode switches in the other filters.

The secondary of T104 is resonated by the C120-C122 series combination, and they also act as an impedance transformation voltage divider. T105 and C121 form a series resonant circuit, and its output is coupled to C124-C125, which again transforms the impedance of the signal line, and also resonates with output transformer T106. The combination of the two parallel resonant circuits and one series resonant circuit forms a bandpass filter of excellent characteristics. Output switching is provided by R118, in conjunction with CR113, in a manner identical to the input switching scheme.

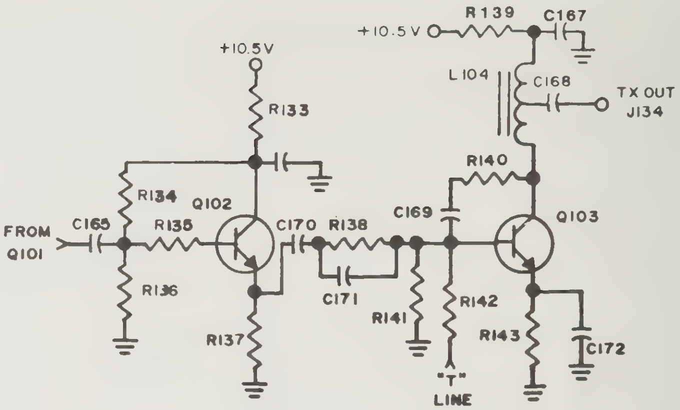
The output of the preselector filters is coupled to the preamplifier stage, Q101, via autotransformer T119 and coupling capacitor C163. R127 provides a DC path to ground for the switching diode, while C159 effectively grounds the bottom of T119 for the signal.



Bias for Q101 is provided by both the source circuitry and the R128-R129 combination. So far as gate two is concerned, the bias is determined by the source network and what ALC is applied to it from the ALC circuitry, to be discussed later. CR122 and CR123, biased in the forward direction by the voltage applied to R150, provide a fixed drop of approximately 1.4 volts, while additional bias is supplied by the drop across source resistor R132. C160 and C161 are both bypasses.

Output, via C165, is taken from Q101 and applied to transmit preamplifier Q102, which is a conventionally biased emitter follower stage. This in turn is coupled via C170 to frequency-equalizing network R138-C171, and then to the base of feedback amplifier Q103. Gain of this stage is determined by the ratio of R140 to the impedance of the R138-C171 combination. Bias is applied only in the transmit condition, from the "T" line, via voltage divider R141-R142. The emitter circuit is composed of R143 bypassed by C172.

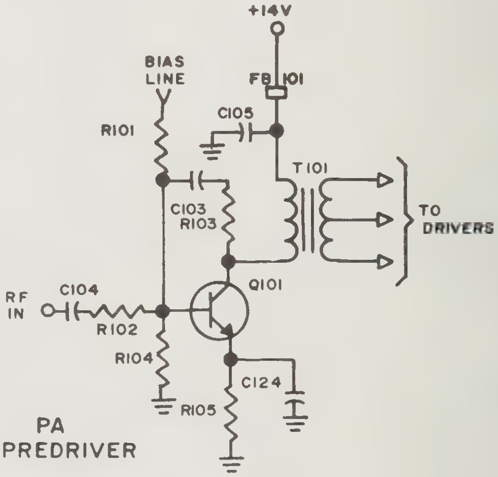
Output is taken from a tap on autotransformer L104 and coupled through C168, providing a low impedance signal that can be coupled through a coaxial cable to the power amplifier board with minimum loss. C167 and R139 form a supply line decoupling network.



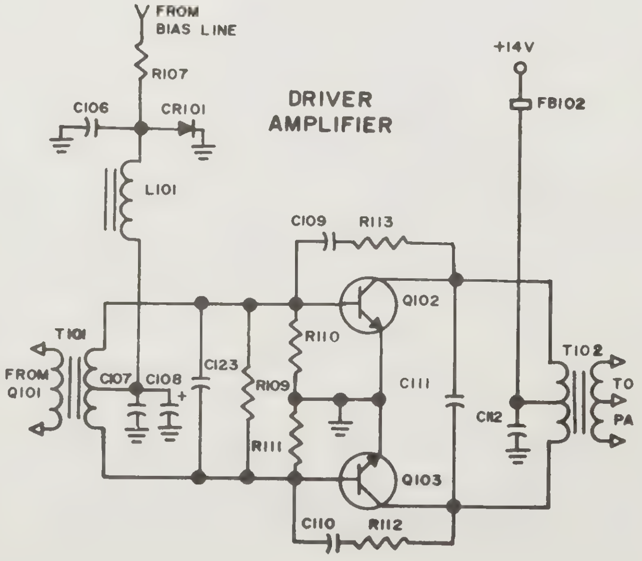
TRANSMIT PREAMP

POWER AMPLIFIER

From the preselector, the transmit signal is applied to PA predriver Q101 on the power amplifier board via coupling capacitor C104. Since Q101 is a feedback stage, its gain is set by the ratio of feedback resistor R103 to input resistor R102, while bias is taken from the PA bias line via voltage divider R101-R104. The emitter circuit is conventional, and the output is taken via T101, which provides a push-pull signal for the following driver stages. C105 and ferrite bead FB101 provide decoupling for the collector supply line.



PA PREDRIVER



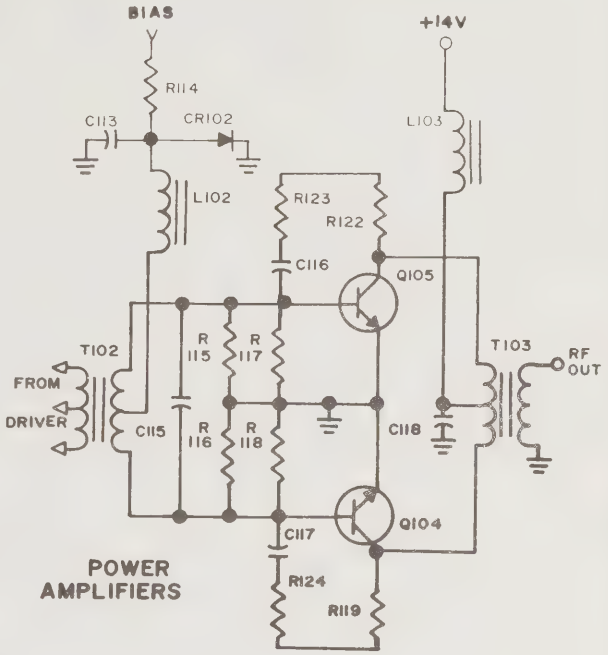
DRIVER AMPLIFIER

The driver amplifiers, fed by transformer T101, are supplied bias via the center tap of that transformer and RF choke L101. R107 and CR101 form a forward-biased diode regulator circuit to control the amount of bias supplied. C106, C107 and C108 are bypasses.

The C123, R109, R110 and R111 network assures the correct load impedance on the secondary of T101. The R113-C109 and R112-C110 networks provide feedback for the stages. C111 helps to provide proper frequency response at the primary of the interstage transformer, T102, while C112 and FB102 provide required decoupling of the supply line.

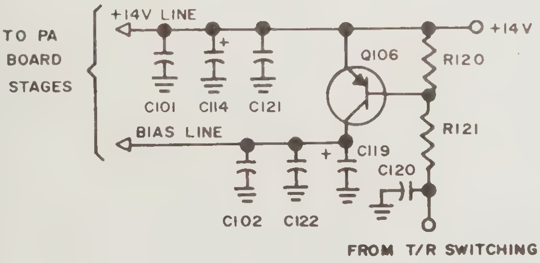
Power amplifier stages Q104 and Q105 both are also driven in push-pull, from the secondary of transformer T102. Bias is supplied in a manner similar to that of the driver stages, and again, a network composed of C115, R115, R116, R117 and R118 assure proper termination impedance for the secondary of T102. Feedback networks are provided for these stages also, and output appears across the secondary of T103. C118 and RFC L103 provide supply decoupling.

The bias line for the PA board is controlled by Q106. Normally, the base of this device is held at the +14 volt supply potential through R120, thus cutting it off. When, however, the transmit mode is activated, the T/R switching circuits ground the lower end of R121, thus lowering the voltage on the base of Q106 and allowing it to saturate. This energizes the bias line and all circuits supplied from it.



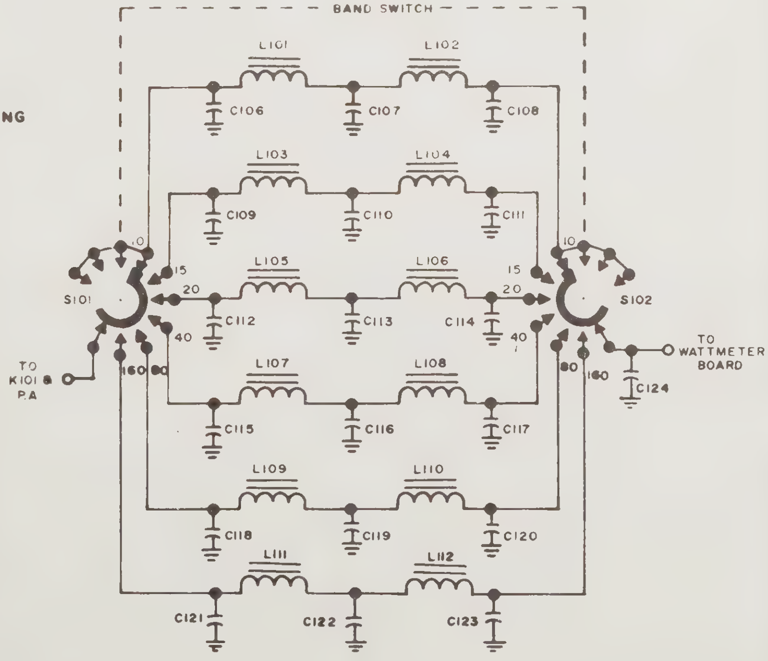
POWER AMPLIFIERS

Capacitors C102, C122 and C119 on the bias line, and C101, C114 and C121 on the +14 volt supply line assure good bypassing at all frequencies from audio to radio.



PA BOARD BIAS

The output from the power amplifier is coupled to switch S101 on the bandswitch/low pass filter board, and thence through the appropriate low pass filter for the band selected, and out through S102 to the wattmeter/SWR board (covered later). Each of the low pass filters on this board is a five pole type, designed to greatly attenuate out-of-band harmonics and spurious outputs.



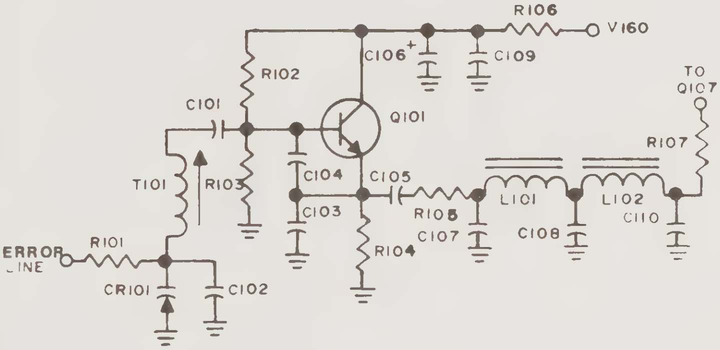
Oscillator Theory

Band Voltage Controlled Oscillators.......5-1
Permeability Tuned Oscillators (PTOs).....5-3
PTO Switching.............................5-4
Local Oscillator Mixer....................5-5
Carrier Oscillator........................5-7
CW Sidetone Oscillator....................5-7
-5 Volt Oscillator/Regulator..............5-8

BAND VOLTAGE CONTROLLED OSCILLATORS

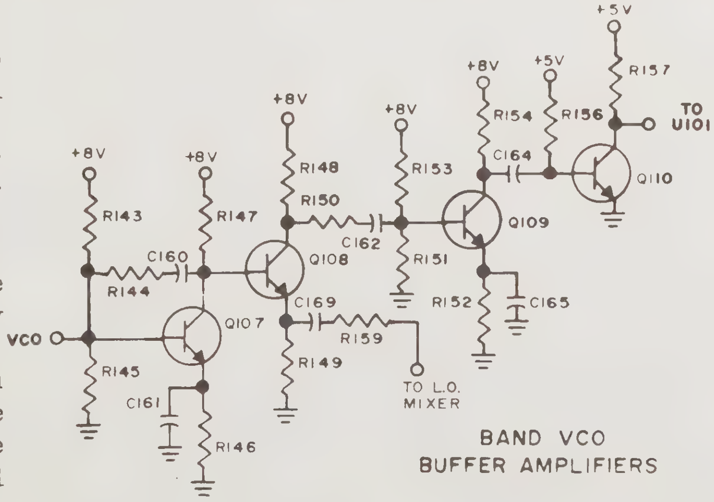
Each of the six bands has its own voltage controlled oscillator (VCO) that operates at a frequency 14.5 MHz higher than the low edge of the band itself. Each of these VCOs is the same as the other, with the exception of the values of the frequency sensitive components, so only one will be discussed here, that for the 160 meter band.

The oscillator circuit is derived from the basic Colpitts oscillator, with C103 and C104 providing the tapped capacitance for feedback. T101 and the parallel combination of varicap CR101 and C102 provide the tuned circuit. Voltage applied by the error line through isolation resistor R101 tunes the varicap. C101 is a blocking capacitor, and bias is developed by voltage divider R102-R103. Output is taken through blocking capacitor C105 and series resistor R105, and is filtered by a low pass filter composed of inductors L101 and L102, with C107, C108, and C110.



BAND VCO

The output of the filter is coupled through series resistor R107 to a line common with the other five oscillators, and to the base of summing amplifier Q107. R144 and C160 provide a feedback path from collector to base of this stage, so the gain of Q107 is determined by the ratio of R144 to R107. (See the "Servicing Hints" section of this manual for a discussion of feedback amplifiers). Bias for Q107 is conventional, as is the emitter circuitry. The collector of Q107 is directly coupled to the base of Q108, making them into a high-gain compound pair. Output to the L.O. mixer is taken from the emitter of Q108 through C169 and R159, while an additional output is taken from the collector through R150 and C162, and applied to the base of amplifier Q109. This stage then drives level translator Q110 through C164, and bias is applied through dropping resistor R156. The output of Q110, now a square wave at a TTL-compatable level, is directly coupled to the prescaler of the band phase locked loop. (See the "Servicing Hints" section for a discussion of the PLL.)



BAND VCO
BUFFER AMPLIFIERS

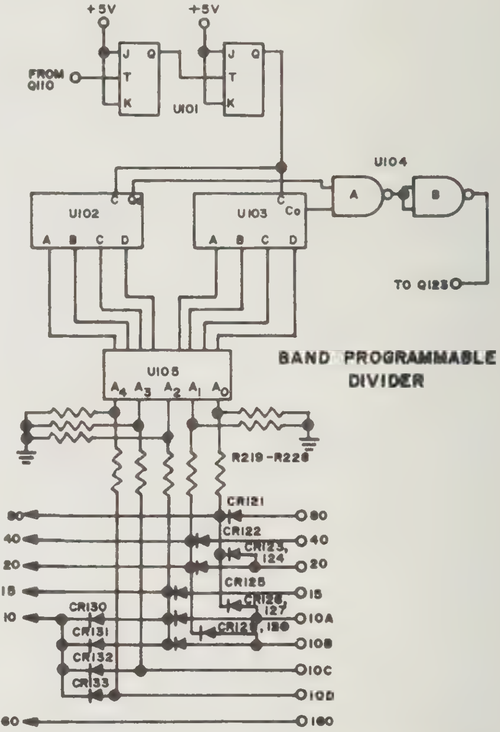
U101, a dual J-K flip flop, is used as a divide-by-four prescaler for the divider chain. Its output is coupled to the clock inputs of U102 and U103 in parallel. The outputs of U102 and U103 are gated through U104A to produce an output pulse when the counter has completed its counting cycle; this pulse is then inverted through U104B, and used to drive the phase detector driver, Q123.

Programming of counters U102 and U103 is accomplished by the diode matrix composed of CR121 through CR129, and memory U105. For 160 meter operation, all five input lines to U105 are low, or binary 00000; for 80 meters, the coding is 00001; for 40, 00010; for 20, 00011; for 15, 00100; for 10A, 00101; for 10B, 00110; for 10C, 01000; and for 10D, 10000. Band program memory U105 decodes these addresses into the bit combinations necessary to set the required division ratios into counters U102 and U103, as shown in the accompanying memory map of U105.

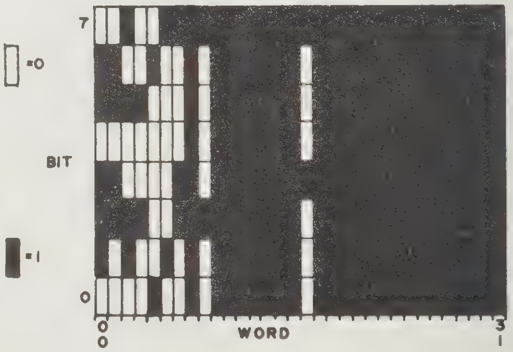
Additional diodes (CR130 through CR133) are connected to the 10 meter control lines which provide a common 10 meter bus, regardless of which segment of the band is selected by the BAND switch, and is used to energize the 10 meter VCO and 10 meter L.O. filter.

The output of the programmable divider is coupled through R232 to the base of driver Q123. Its output is coupled to the phase comparator, U109. The other input to the phase comparator is the 125 kHz signal from U101B, the loop reference oscillator and divider. Y101 controls the frequency of the oscillator at 2 MHz, with C246 providing a means of adjusting the calibration.

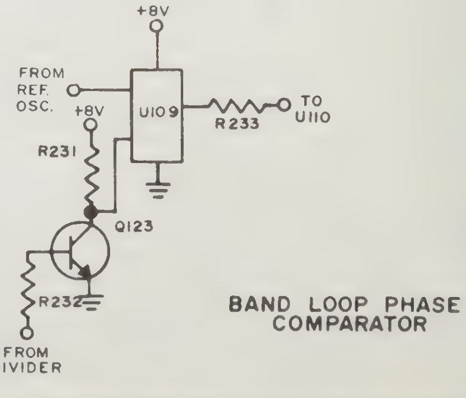
Output of the phase comparator, U109, is coupled through R233 to the inputs of two amplifier stages, U110A and U110B. Capacitor and resistor combinations C248-R233 and R239-C249 form the loop filter. For 160 through 15 meter operation, U110A is used as the error amplifier, with its output driving the varicaps in those band oscillators; for the 10 meter band, U101B, with higher gain, is used to drive the varicap of that band's oscillator. More gain is required be-



BAND PROGRAMMABLE DIVIDER



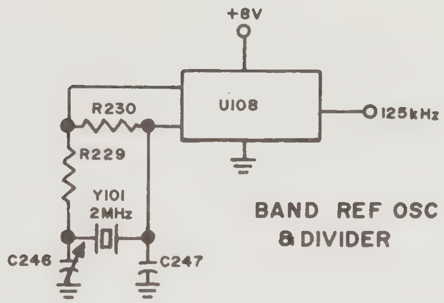
U105 MEMORY MAP



BAND LOOP PHASE COMPARATOR

cause the higher divider numbers also reduce the loop gain.

In this manner, the band VCOs are constrained to oscillate on precise multiples of the crystal controlled reference frequency of 125 kHz. The exact multiple is the total division performed on the VCO frequency, including that of the prescaler.

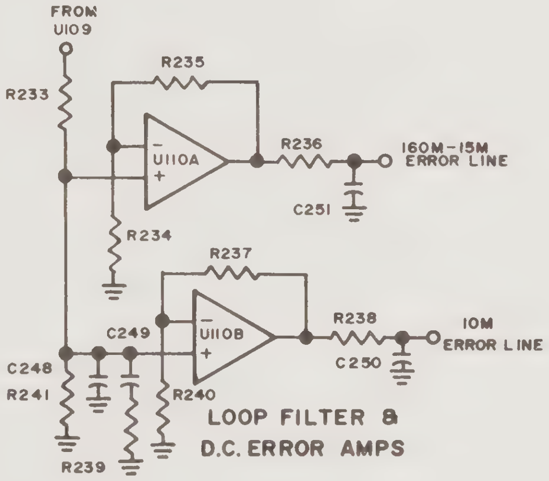


BAND REF OSC
& DIVIDER

PERMEABILITY TUNED OSCILLATORS

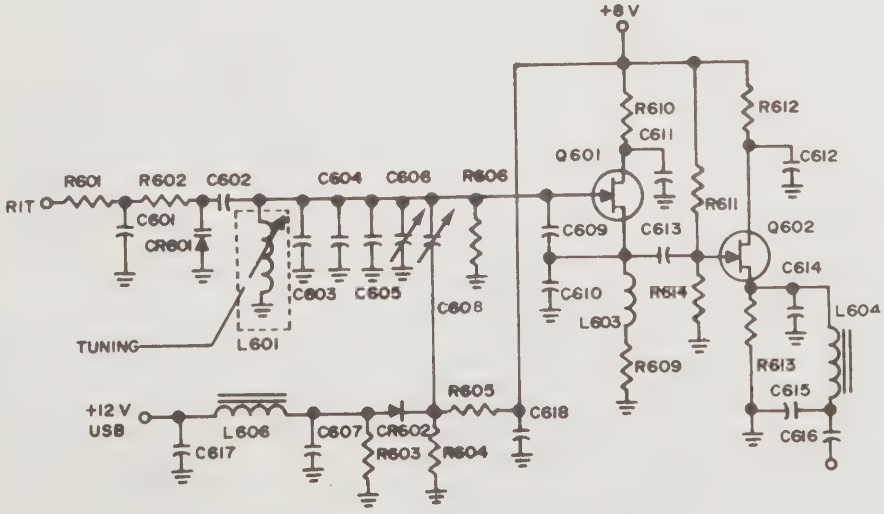
There are two identical permeability tuned oscillators in the ASTRO 102BX. Only one is described here.

The permeability tuned oscillator is the primary means of varying the frequency of the transceiver. It is a Colpitts circuit, and the tuned frequency is determined by L601 in parallel with the appropriate capacitors, CR601, C603, C604, C605 and C606. The frequency tuned is 5.0 to 5.5 MHz, and this combination of capacitors is chosen to provide temperature compensation for low drift, and must be replaced with exactly the same types to maintain this stability, if replacement becomes necessary.



LOOP FILTER &
D.C. ERROR AMPS

R606 provides a DC return for the gate bias of Q601, which is developed by R609. Feedback to sustain oscillation is achieved by the C609-C610 divider combination. The drain is decoupled to ground by R610 and C611, while the output is taken from the source to drive FET buffer Q602.



PERMABILITY TUNED OSCILLATOR

The bias generated by source resistor R613 of Q602, whose value is set by the desired output impedance, is greater than is desirable for this FET; hence, a positive voltage to offset this is developed by the R611-R614 combination. Output is taken via low pass filter C614-L604-C615 to assure good spectral purity.

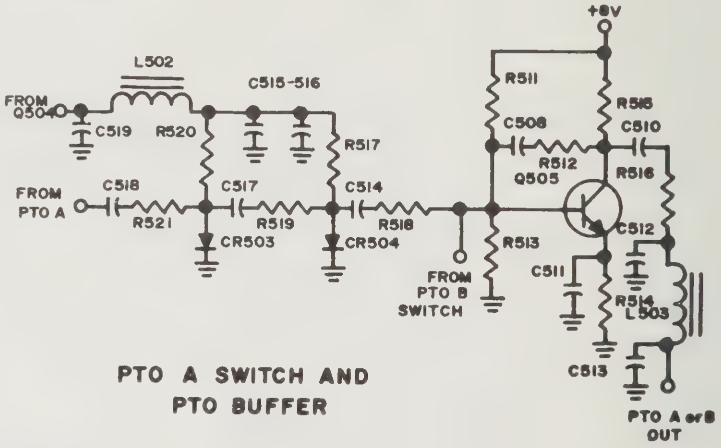
When operating in the USB mode, the required frequency shift is set by C608, which is switched in parallel with the tuned circuit by CR602.

When RIT is used, the required shift is obtained by varactor diode CR-601, connected in parallel with the tuned circuit by blocking capacitor C602. The DC tuning voltage for this varactor is applied through the R601-R602 combination, and is decoupled by C601. C606 is a calibration control.

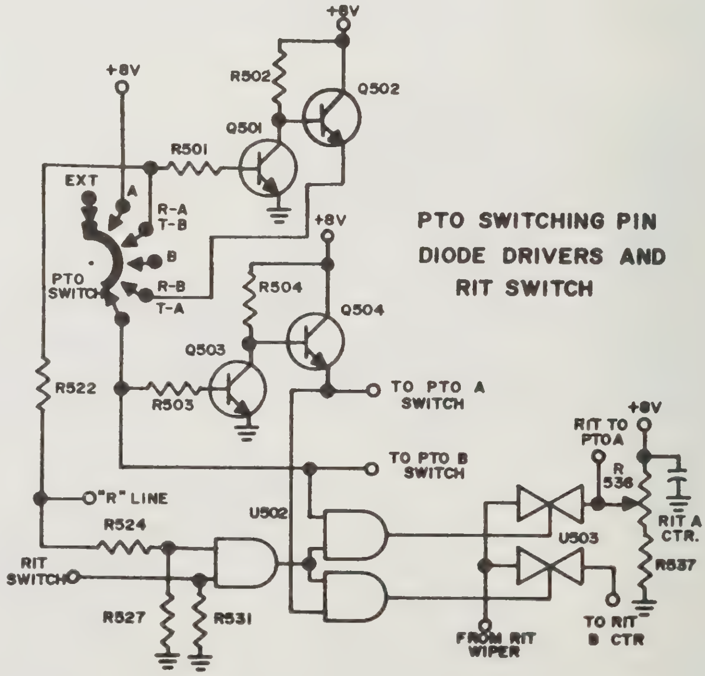
PTO SWITCHING

The switches are used for the PTO outputs; only one will be described here, as both are identical and the description of one will apply equally to both.

A control voltage, the generation of which will be discussed later, is applied to decoupling RF choke L502 from Q504. C519, C515 and C516 are bypass capacitors. This voltage is applied to PIN diode attenuators CR503 and CR504 through isolating resistors R520 and R517, causing the diodes to present a low resistance to ground. Since the signal from the PTO passes through R521 and R519 (as well as blocking capacitors C518 and C517, it is dropped by these two resistors, with essentially none passing through to the base of the buffer, Q505. Thus, when this control voltage is removed, the PTO output is allowed to pass to the buffer. Q505 is a summing amplifier, with its gain determined by the ratio of R512 to R518 (and others, when the other section is in use). Bias and emitter circuitry are conventional, and output is taken from the collector through low pass filter C512-L503-C513 to assure that no harmonics or spurs are passed on to



PTO A SWITCH AND PTO BUFFER



PTO SWITCHING PIN DIODE DRIVERS AND RIT SWITCH the mixer. In the off condition, the switch provides greater than 100 dB of attenuation to the unwanted PTO output.

The control signal for the PTO switching circuitry is derived from the front panel PTO switch. In the PTO A position, +8 volts is applied to the PTO B switching circuitry directly through the switch contacts, thus disabling PTO B. It is also applied to the base of Q503, causing it to saturate; since Q504 is direct coupled to its collector, the latter is then turned off and no voltage appears on the PTO A control line, allowing its signal to pass and appear at the output.

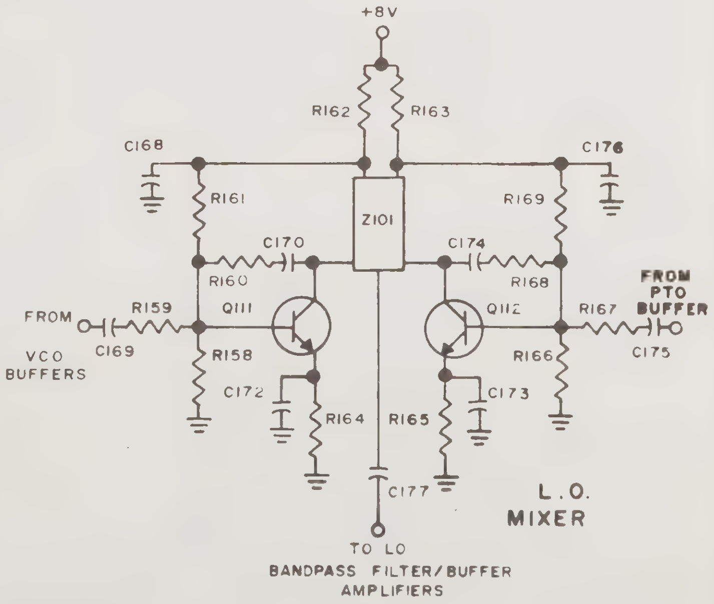
In the PTO B position, no voltage is applied to the PTO B line, allowing it to function, and Q504 is turned on, allowing voltage to appear on the PTO A control line, disabling PTO A.

For split operation, "R" line voltage is used (through the PTO switch on the front panel) to drive the PTO B control line and the Q503-Q504 combination. Thus, which PTO is turned on depends upon whether the "R" line is high or low (receiver on or off). For reverse split operation Q501 and Q502 provide an inverted "R" line voltage (+ for transmit, 0 for receive) which is used instead.

When the RIT switch on the front panel is on, and the transceiver is in the receive condition ("R" line high), AND gate U502A is on, and its output is high. This high is coupled to one input of U502B and U502C. The other inputs of each of these AND gates is connected to the PTO A and PTO B control lines, respectively. Thus, when the PTO B control line is high (and the PTO A line is low), analog switch U503A is turned on and the voltage from the RIT control wiper is applied to the RIT input of PTO A. R536 and R537 provide center adjustment for the RIT circuitry in the PTO.

When the PTO A control line is high, the remaining section of the AND gate is on, turning on U503B, and supplying the RIT control voltage to PTO B. A center adjustment similar to that for the A PTO is included in the circuit although not shown in the diagram, for simplicity.

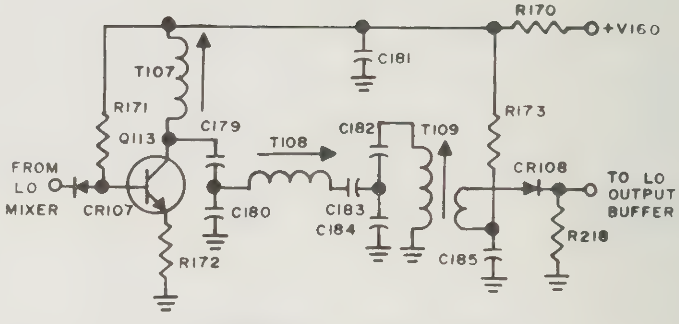
LOCAL OSCILLATOR MIXER



Output from the band synthesizer (VCO buffer) is applied to mixer driver Q111 through blocking capacitor C169. Q111 is a feedback amplifier whose gain depends on the ratio of R160 to R159, and its output is applied to one input of the L.O. mixer Z101, which is a doubly-balanced mixer identical to others used in this transceiver. Also, the output of the selected PTO is applied to Q112 through blocking capacitor C175. This stage too, is a feedback amplifier, with its gain determined by R167 and R168. Its output is similarly coupled to Z101.

Output of Z101 is at the operating frequency plus 9 MHz, and is processed through one of six bandpass filter-amplifier combinations. Since all six are identical in operation, only the 160 meter version will be discussed here.

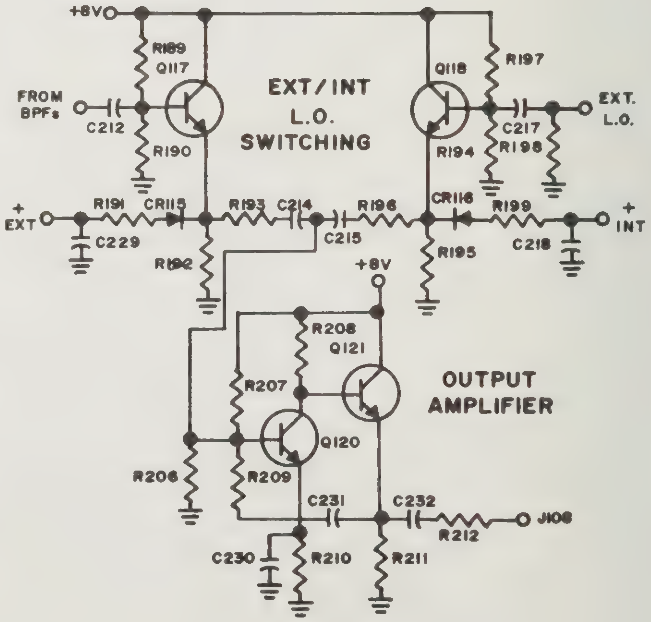
Collector and switching power comes from the 160 meter control line. R171 supplies bias for amplifier Q113, and also biases switch CR107 on; the voltage drop across its ground return resistor (not shown in the simplified diagram) reverse biases similar diode switches in the other bandpass filter circuits.



LO BANDPASS FILTER/BUFFER

Q113 operates without an emitter bypass capacitor, and is therefore a bit degenerative; its collector is composed of a parallel resonant circuit formed by T107 and the C179-C180 combination. Output is taken from the low impedance tap between these two capacitors, and coupled through series resonant circuit T108-C183, then stepped back up to higher impedance by the output parallel resonant circuit T109-C182-C184. Output switching is via CR108, biased on through R173, the secondary of T109, and R218. C185 is a bypass for the signal. Again, the voltage drop across R218 reverse biases the output switching diodes in the other filters.

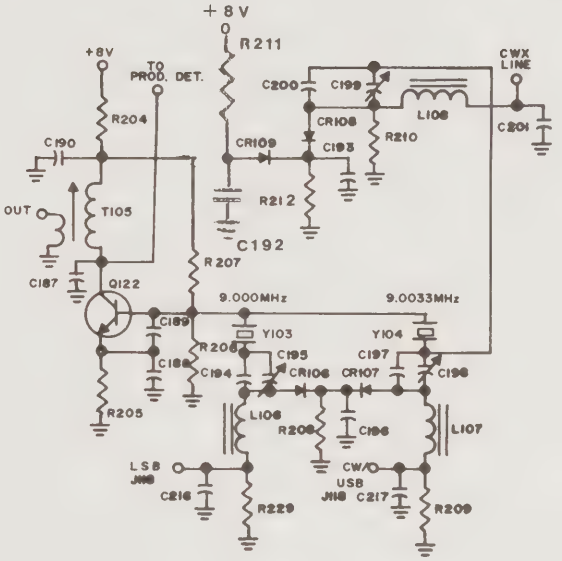
When the PTO switch on the front panel is in the EXTernal position, a voltage is applied through R191 to CR115, and thence to the emitter of L.O. buffer Q117. This reverse biases this stage's base—emitter junction, preventing internally-generated L.O. energy from being passed on. At the same time the voltage drop across R195 reverse biases CR116 (since the INT line is low) and Q118 acts as a buffer for externally supplied L.O. signal, and it is passed on through R196 and C215 to the L.O. output amplifier, Q120.



When the PTO switch is in any other than the EXT position, the INT and EXT voltages are reversed, and internally generated L.O. signal is passed on to Q120.

Q120 and Q121 are directly coupled, and amplify the L.O. signal selected by the switching circuit. The emitter of Q120 is wired conventionally, while R206 and R207 are the bias divider. Feedback from the emitter of Q121 to the base of Q120 is provided by the R209-C231 network. Output is taken from the emitter of Q121 through C232 and R212, and routed to the exciter board.

CARRIER OSCILLATOR



CARRIER OSCILLATOR

Since the ASTRO 102BX utilizes only one crystal filter, the carrier oscillator frequency must be shifted appropriately when selecting USB, LSB or CW modes, so that the signal will fall within the pass band of the filter. Only one oscillator is used, with a crystal switching scheme used to select the proper crystal, depending upon the mode selected.

Q122 is the oscillator, with bias provided by R206 and R207. Feedback is provided by C188 and C-189. R205 is the emitter resistor. The collector load is comprised of T105 and C187, which form a parallel resonant circuit. R204 and C-190 form a decoupling network.

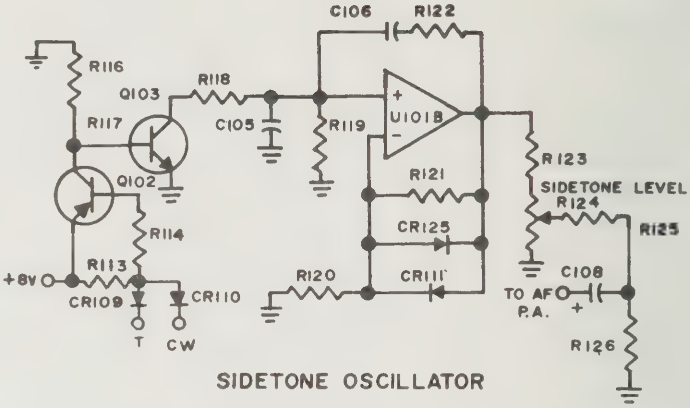
In the LSB mode, a + voltage is applied through L106 which forward biases diode CR106, with R208 serving as a ground return. In this manner, C196 effectively grounds the lower end of C194 and C195, placing Y103 into the circuit between Q122's base and ground; the oscillator then operates at Y103's frequency, as trimmed by C195.

At the same time, the voltage drop across R208 reverse biases CR107, preventing Y104 from having any effect on the oscillator frequency. In the USB or CW receive modes, the + voltage is applied via L107 to CR107, forward biasing it in the same manner as CR106 in the LSB mode. CR106 is now reverse biased, and Y104 controls the frequency of oscillation. The exact frequency is set by C198.

When transmitting in the CW mode, an offset is desired to place the signal into the passband of the IF filter. In the ASTRO 102BX, this offset is approximately 800 Hz from the receive condition, and is provided by "pulling" the frequency of Y104. In transmit, the CWX line is high, forward biasing CR108 and reverse biasing CR109. C199 and C200 now affect the frequency of Y104, and C199 is adjusted during alignment to provide the desired frequency offset.

CW SIDETONE OSCILLATOR

When operating in the CW mode, a sidetone is provided in the transceiver's loudspeaker so that the operator may monitor his own sending. This signal is generated by CW sidetone oscillator U101B. The frequency of the sidetone is determined by the positive feedback path for U101B, that is, C-106, C105, and R118. The gain of the stage is determined by the inverting feedback path, R121, CR125, CR111 and R120, with the diodes providing a leveling action for the oscillator output. The output is applied to sidetone level control R124, and thence to the audio frequency power amplifier via coupling capacitor C108.



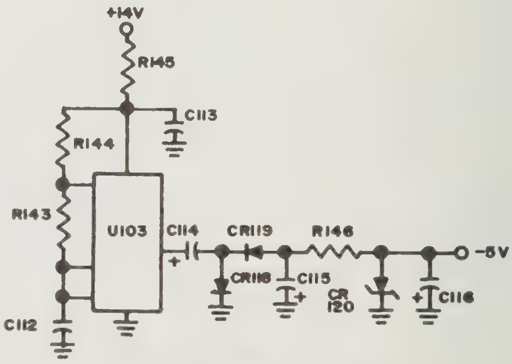
SIDETONE OSCILLATOR

Keying of the sidetone oscillator is accomplished by Q102 and Q103. When in the CW mode, CR110 is reverse biased, but the base of Q102 is held low by the "T" line, maintaining Q102 in a conducting state. When the key is closed, the "T" line goes high, allowing the base of Q1o2 to rise, and cutting off conduction in R116 in the collector circuit

Thus, when the key is closed, and the MODE switch is in the CW position the base of Q103 is low due to the fact that Q102 is not conducting, and it does not conduct. This allows U101B to oscillate. When the key is opened, the "T" line goes low, CR109 conducts, and the base of Q102 is grounded. It thus conducts, creating a voltage drop across R116, and biases Q103 on. This prevents U101B from oscillating.

-5 VOLT OSCILLATOR/REGULATOR

The negative five volts DC required by various portions of the transceiver circuitry is provided by oscillator U-103, a timer IC connected in the astable mode. The frequency of oscillation is set by the R143, R144 and C112 combination. Output is taken through capacitor C114, and applied to a voltage doubler composed of CR118 and CR119. C114 and C115 are an integral part of the doubler circuit, charging alternately on different half-cycles of the oscillator output. Their individual voltages add during discharge, providing the required voltage doubler action.



—5V OSCILLATOR AND REGULATOR

The negative voltage thus generated is regulated by the combination of dropping resistor R146 and zener diode CR120, in a conventional zener regulator circuit. Output is filtered by C116.

Ancillary Circuits

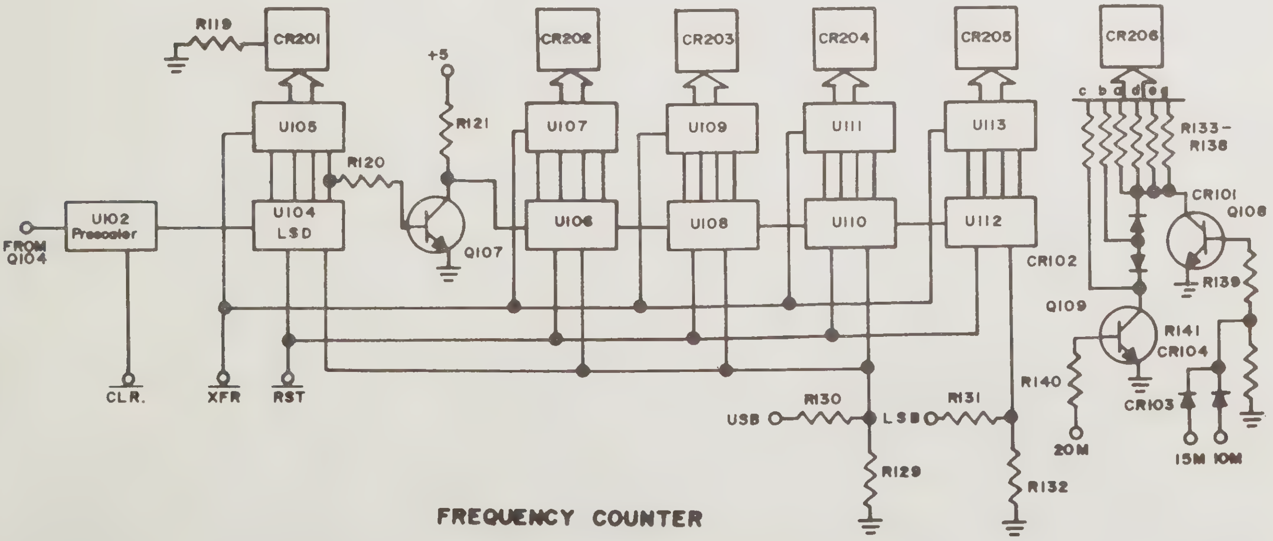
Frequency Counter/Digital Dial......................6-1
10.5 Volt Regulator.................................6-4
Wattmeter/SWR Detector..............................6-4
Automatic Level Control (ALC).......................6-5
VOX...6-6
Keying Control......................................6-7
CW Timer..6-8
Noise Blanker.......................................6-9
Passband Tuning Indicator...........................6-10

FREQUENCY COUNTER

The ASTRO 102BX uses an accurate, crystal controlled time base and a frequency counter to provide the operator with a digital display of the transceiver's operating frequency. This counter measures the local oscillator frequency, corrects for the fact that this is 9 MHz above that which the transceiver is receiving or transmitting, and displays the results.

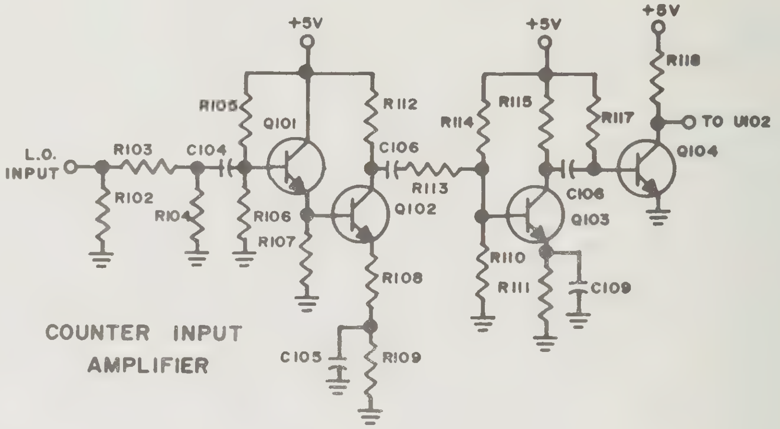
The incoming local oscillator signal, taken from the emitter of Q114 on the exciter board, is amplified and processed by the counter input amplifier. R102, R103 and R104 form an attenuator network, used to reduce any spurious signals generated in the counter, preventing them from being coupled back into the receiver circuitry. The overall design of the counter includes extensive planning to prevent "counter birdies"--an unfortunate phenomenon in all digital dial radios. This includes the solid shielding box and cover that the counter is enclosed in, feedthrough capacitors on leads penetrating this shield, high isolation amplifiers, input attenuator, etc. The results are that the unavoidable spurs are reduced to insignificant levels. The technician is cautioned to exercise particular care in maintaining the integrety of this shielding.

C104 couples the signal to the base of the first of two devices connected as a compound pair, Q101 and Q102. Bias for Q101 is provided by conventional voltage divider R105-R106, and the drop across its emitter resistor, R107, provides bias for Q102. The emitter resistance of Q102 is divided into two parts, R108 and R109, with the latter bypassed by C105. This arrangement compensates for the inherent 6 dB-per-octave rolloff in transistor gain, thus levelling its response over the range desired. Output from Q102 is taken via the collector, coupled through C106 and R113, and applied to the base of Q103. This stage is conventional in all respects, with R114



FREQUENCY COUNTER

and R110 forming the bias network and R111, bypassed by C109, the emitter circuit. The output of Q103, now gain-levelled and amplified, is applied to the base of Q104 through C108. Since the desired output of this stage is a square wave suitable for driving the counter prescaler, and linearity is no longer of importance, R117 suffices to provide bias for Q104. The collector is directly coupled to the input of the prescaler.



COUNTER INPUT AMPLIFIER

U102, a divide-by-ten counter, reduces the frequency of the incoming signal from the input range of approximately 10.8 to 39 MHz (9 MHz above the transceiver operating frequency) to 1.08 to 3.9 MHz for the input of the counter proper. It is cleared at the end of each count cycle by the negative-going pulse on the clr line. This line provides the necessary gate function to drive the counter proper.

The counter chain is conventional, with two exceptions; the output of the prescaler is still too high for reliable operation of the CMOS devices, particularly complex devices such as counters, on a 5 volt supply; hence, in this decade, a low-power Schottky device is used. This counter, the least significant figure decade (100 Hz on the display), counts using the conventional BCD code, and its four output lines drive U105, a combined storage latch, decoder, and seven-segment display driver. When the end of the count cycle is reached, the xfr line is brought momentarily low causing the latches in U105 to accept the data being fed to it by U104, and to store it until updated at the end of the next counting cycle. This data is then decoded, and the proper segments of the display, CR201, are activated, displaying the count. CR201 is a common anode type of display, so grounding the input pins for the appropriate segments will cause them to light. Current is limited to the correct value (15 mA) by internal circuitry of U105. R119 is used to provide a path to ground for the decimal point of CR201, and to limit its current, so that it is lighted at all times the radio is on.

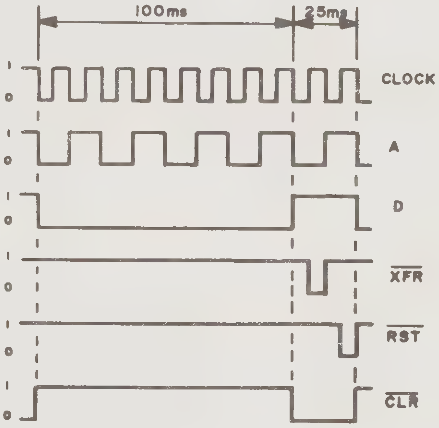
The second somewhat unusual feature of the counter is that the output of the LSD decade is not taken from the counter "carry" output, for the pulse presented here is much too narrow for the following, CMOS decade. Instead, output is taken from the Q_D terminal, through current limiting resistor R120, and applied to the base of Q107. The collector, with pull up resistor R121, is connected to the input of the next decade counter, U106. The pulse thus obtained is much wider than that presented at the "carry" output, and is inverted to the correct polarity.

The subsequent decades of the counter are all of the CMOS family--U106, U108, U110 and U112. Each has a "carry" output which is used to drive directly the "up count" input of the subsequent decade. Each has four BCD coded data output lines which are connected to the latch/decoder/driver IC corresponding to that decade, i.e., U107, U109, U111 and U113. Each of these in turn operates the appropriate display indicators. The latches are all strobed once per count cycle to update the display, in the same manner as described for U105.

In order to read the incoming frequency correctly, to the nearest 100 Hz, the counter would have to be gated "on" for 10 milliseconds; however, since we use a prescaler with a divide-by-ten ratio, we must extend the count cycle time to 100ms. This is accomplished by the clr line of U102, the prescaler. It operates, and passes on its output pulses, only when that line is high; thus, for the time it is low, the counters stop, but store their present count while the latches are strobed, thus passing that count to them (and the displays), and then are preset by the rst line, ready to start the next count cycle.

Unfortunately, it takes a finite time to (a) stop the counting, (b) transfer the count to the latches, (c) clear the counters and (d) start the next count cycle. This is accomplished, as shown in the accompanying counter timing chart, during the 25 ms immediately following the counting portion of the counting cycle. This overall cycle translates to a rate of eight updates of the display per second.

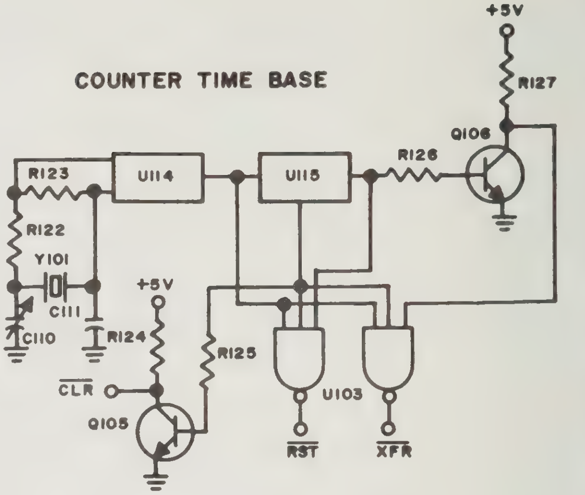
COUNTER TIMING CHART



For a laboratory-style frequency counter each decade would be reset to zero (BCD 0000) at the end of each complete count cycle; but here, we have to account for the fact that the frequency being counted is 9 MHz above the operating frequency, and for the fact that we have different offsets (carrier oscillator frequencies) for USB and LSB operation. This makes it necessary to preset the counters to a number different than zero in order to compensate. When operating in the USB mode, the number to which the counters must be preset are (MSD to LSD) 09967; in the LSB mode, it is 10000. This is accomplished by hard-wiring the USB and LSB lines to the appropriate load inputs of the counter chips, and using the rst line to drive the "load" control pins. Then, when the active counting portion of the count cycle begins again, each counter advances through zero (overflow) to a value appropriate for the transceiver operating frequency, and this value is displayed. Any carry from the MSD counter is discarded and ignored.

Another way of stating the matter is that the counters are preset to the above numbers, which correspond to a number that is 9,000.0 (or 9,003.3) subtracted from the overflow value of the counter chain, i.e., 10000.0. This is, in fact, a negative number, although the counter does not recognize it as such---in this way, the counter subtracts the carrier oscillator frequency from the L.O. frequency, as is required for a correct display.

It is not necessary for the 10 MHz digit to be arrived at by a counter, and to do so would add unnecessary complexity to the circuit. Instead, it is hard-wired such that, when the BAND switch is on the 160 through 40 meter positions, it is extinguished; when it is in the 20 meter position, Q109 is saturated and segments B and C of the display are grounded through appropriate current limiting resistors, and the numeral "1" is therefore displayed; and in the 10 and 15 meter positions of the BAND switch, Q108 is saturated, grounding the current limiting resistors for segments A, B, D, E and G, thus displaying the numeral "2".



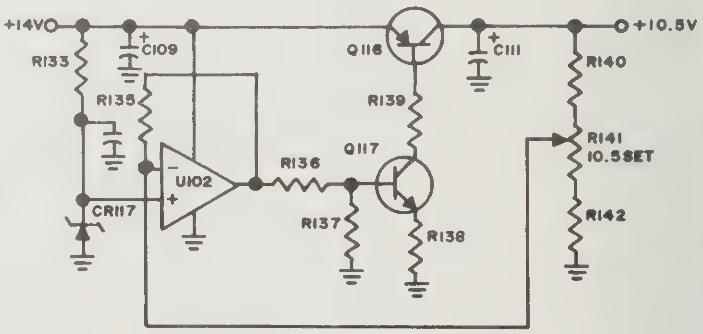
COUNTER TIME BASE

The necessary control signals for the counter ($\overline{\text{clr}}$, $\overline{\text{rst}}$, and $\overline{\text{xfr}}$), are generated by the time base/control circuitry. Crystal Y101, operating at a frequency of 1.31072 MHz, provides the stability necessary for accurate frequency measurement. C110 permits it to be adjusted to precisely the correct frequency. U114 contains the oscillator circuitry, as well as a divider with a factor of 2^{14}, or 16,384. The output of the counter is the square wave signal labelled "clock" on the timing chart. U115 is a decade counter, with outputs taken at the clock-divided-by-two (A) terminal, and the clock-divided-by-ten (D) terminal. The waveforms are shown in proper relationship in the timing diagram. The D output, inverted by Q105, is used as the clr signal; gate U103A generates the rst signal, and gate U103B generates the xfr signal.

10.5 VOLT REGULATOR

A regulated 10.5 volts is required by many of the circuits in the transceiver. It is provided by the circuit comprised of Q116, Q117 and U102. The raw +14 volts is applied to the emitter of PNP series transistor Q116.

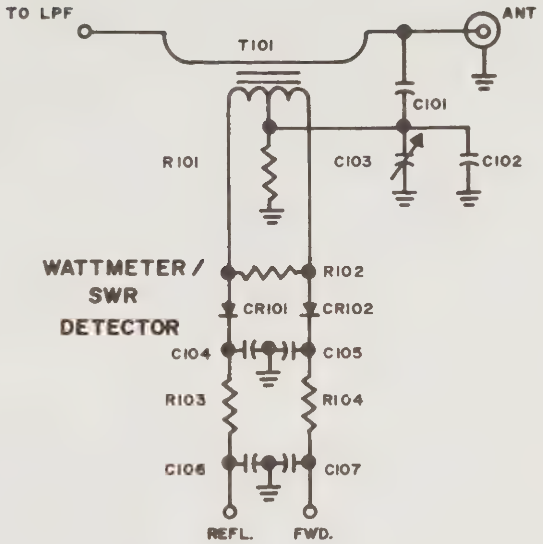
The output of Q116, its collector, is sampled by the R-140-R141-R142 voltage divider, with R141 serving as the "+10.5 set" control, and its wiper is connected to the inverting input of U102. A reference derived from zener diode CR117 through dropping resistor R133 is applied to the operational amplifier's non-inverting input. The output of this stage, a DC voltage, is applied to Q117 and determines the degree to which it is allowed to conduct, and thence its collector voltage. Its collector, in turn, drives the base of Q116 in such a manner as to control its degree of conduction, and thence to maintain its output voltage at the required level.



+10.5V REGULATOR

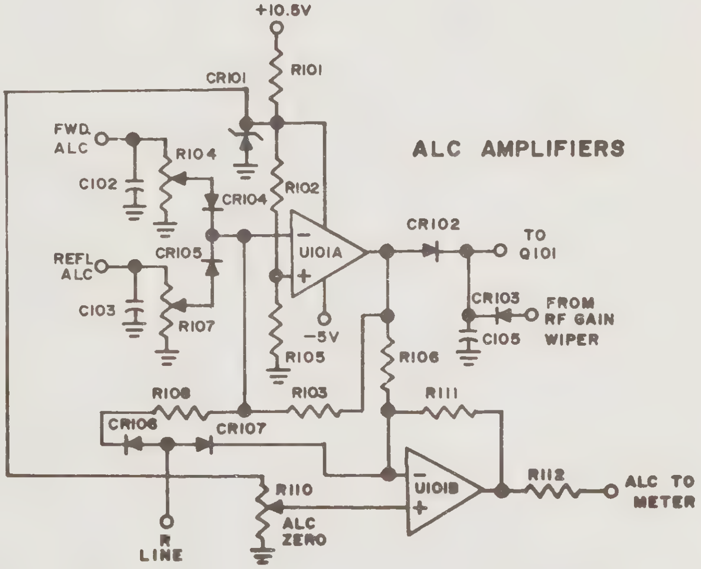
WATTMETER/SWR DETECTOR

The wattmeter board circuitry takes a sample of the current in the antenna line via transformer T101, and a sample of the voltage signal appearing on the antenna line at the output of the transformer via the capacitive voltage divider composed of C101, C102 and trimmer C103, the latter of which allows precise alignment. These signals are combined in such a manner that CR101 produces a rectified output voltage proportional to the magnetude of any signal reflected from the antenna due to mismatch, etc., while CR102 provides a voltage proportional to the power sent to the antenna. These signals are smoothed by the C104-R103-C106 and C105-R104-C107 combinations. These signals are then used in the ALC, SWR, and wattmeter circuitry to provide front panel indications and protective functions for the transmitter power amplifier.



AUTOMATIC LEVEL CONTROL

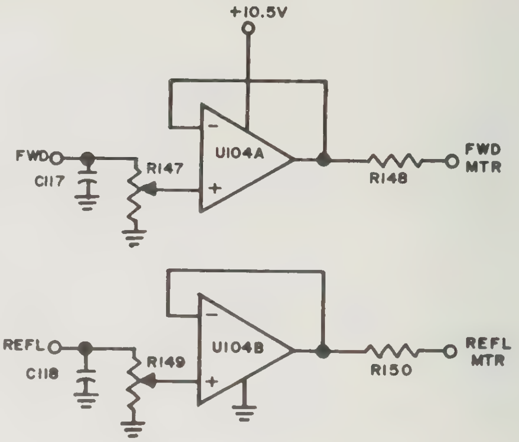
The FWD and REFL signals from the wattmeter board are coupled to the preselector board where they are adjusted by R104 and R107 respectively. C102 and C-103 bypass the lines as they come into the board. CR104 and CR105 act as an "or" gate, allowing the largest of the two signals to pass to the inverting input of ALC amplifier U101A. Feedback for this amplifier is via R103, while R102 and R105 establish a bias at the non-inverting input which sets the quiescent operating point of the output of this stage. R101 and CR101 provide a regulated voltage for this bias network and for U101, as well as for the ALC ZERO adjustment.



If the REFLected ALC signal should exceed a given amount, or if the FWD ALC signal should exceed that produced at the maximum safe output of the PA, CR105 or CR104 will conduct, causing the output of U101A to go more negative. This output will be passed through CR102, and on to Q101 in the preselector, where it is used to reduce the gain of the stage, and thus the output of the transmitter. Manual RF GAIN control is accomplished by inserting a voltage to Q101 via CR103 to the ALC line when in the receive condition. The "R" line, through CR106 and R108, biases the inverting input of U101A in such a manner as to disable it in receive.

The output of U101A is also coupled via R106 to the inverting input of U101B. The ratio of feedback resistor R111 to input resistor R106 determines the gain of the stage, and the output is coupled through limiting resistor R112 to the meter switch, where it may be selected by the operator to give a front panel ALC indication if he so desires. R110, connected to the non-inverting input of U101B, provides a zero set calibration control for the meter reading. When in the receive condition, U101B is also disabled by the "R" line, through CR107, in the same manner as U101A.

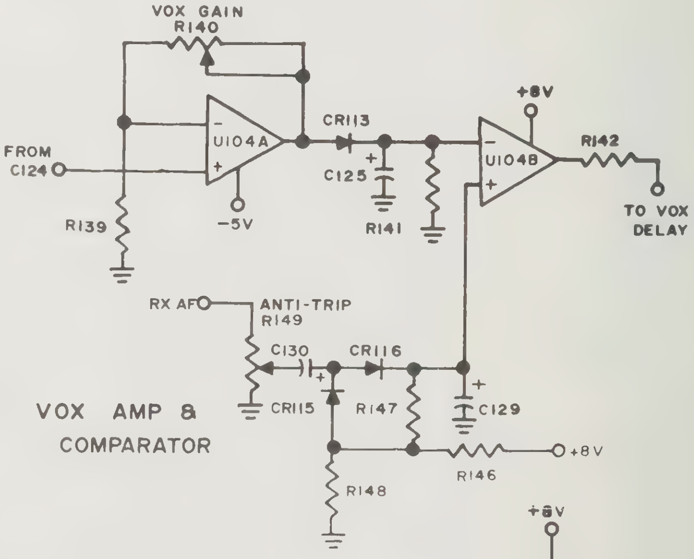
FWD and REFL signals from the wattmeter board are also applied to SWR meter drivers U104A and U104B through calibration potentiometers R147 and R149, respectively. These are unity-gain amplifiers, used only to prevent the meter from loading the wattmeter board output. The operator may select either the FWD or REFL power to be read on the calibrated meter by means of the front panel METER switch. R148 and R150 are current limiting resistors.



SWR METER DRIVERS

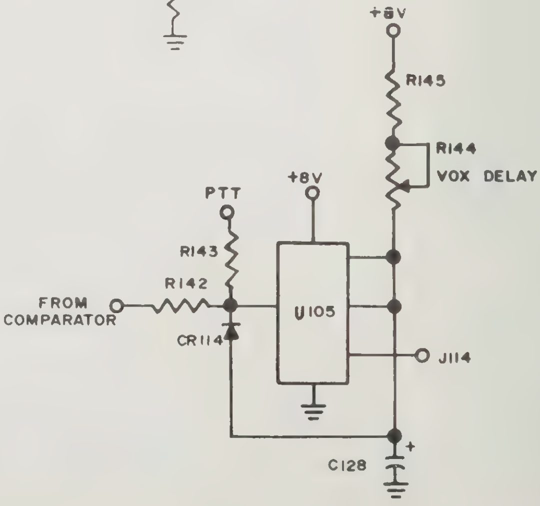
VOX CIRCUIT

When VOX operation is selected by the operator, microphone AF is supplied from the preamplifier to U104A, where it is further amplified. The amount of this amplification is controlled by the VOX gain control, R140. The output of this stage is rectified by CR113, and smoothed by C125, and applied to the inverting input of comparator U104B.



VOX AMP & COMPARATOR

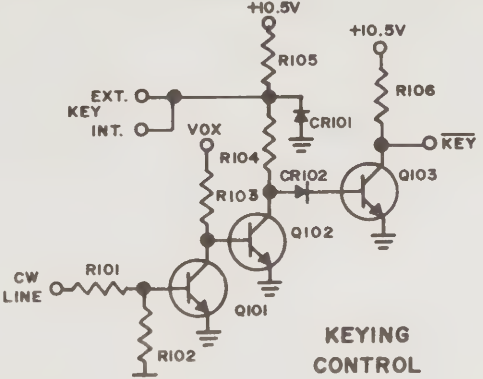
A sample of the received AF, from the speaker line, is applied to the anti-trip control, R149, and then through C130 to CR115 and CR116, which form a voltage doubler rectifier. C129 smooths this DC signal, and it is coupled to the non-inverting input of the comparator. A DC bias, to set the comparator's threshold, is added to the rectified anti-trip signal by the R146-R148-network. The comparator's output will depend upon which of the two input signals are the larger; should the VOX signal be the largest, the output will go low; if the anti-trip signal is larger, it will be prevented from doing so.



VOX DELAY

The output of the comparator is coupled through R142 to the trigger input of timer U105, causing it to activate and its output to go high. This high state is continued for the duration of the low on the input (diode CR-114 prevents C128 from charging during the duration of the low voltage at its cathode) plus the time it takes for C-128 to charge through R144 and R145 to 2/3 of the supply voltage after the low on the input of the timer has ceased. The amount of time it takes for C128 to charge is determined by the setting of R144, thus the amount of delay is controllable. Should VOX not be desired, the VOX/PTT switch connects a + voltage to the timer input via R143 when in the PTT position

KEYING CONTROL

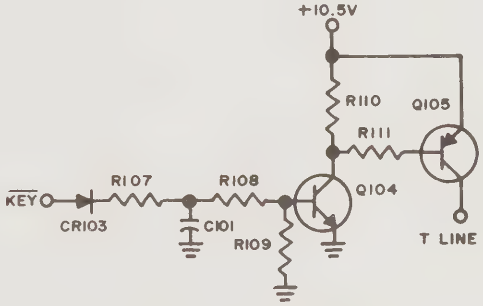


KEYING CONTROL

When in the CW mode of operation, the CW line is high, and it, through R101 biases Q101 on, effectively reducing its collector voltage and cutting off Q102. This makes the collector voltage of Q102 rise, and CR102 conducts in turn, biasing Q103 to an on condition. Thus its collector, the key line, is at a low value, its normal, un-keyed condition.

Should either the EXT KEY or the KEY contacts be closed, the junction of R104 and R105 in the collector circuit of Q102 is grounded, removing the voltage to CR-102 and the base of Q103 causing the collector voltage to Q103 (and the key line) to rise.

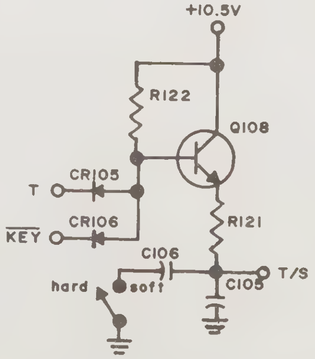
In the USB or LSB modes, the CW line is low, and Q101 does not conduct. PTT operation is then exactly as described above for CW, but if VOX is selected, the VOX circuit provides a voltage to R103 which causes Q102 to saturate, again removing voltage from CR102, and so Q103 ceases to conduct and the key line rises.



"T" LINE CONTROL

The key line, when high, causes CR103 to conduct through R107 and R108 biasing Q104 on. The combination of R107 and C101 control the rise characteristics of the keying waveform, while the falling characteristics are controlled by the discharge of C101 through R108 and the parallel combination of the base-emitter junction of Q104 and R109. When Q104 conducts, its collector goes low, lowering the base potential of PNP transistor Q105, and it saturates, energizing the "T" line.

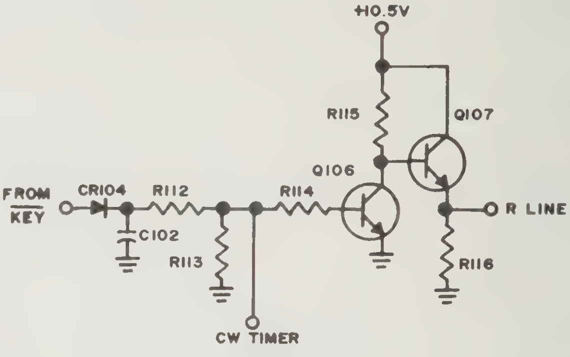
When the "T" line and key line are both high, CR105 and CR106 are reverse biased, allowing Q108 to conduct, and charge C105 through R121. This time constant determines how fast the "T/S" line is allowed to rise to full voltage. The "T/S" line supplies bias to the drivers for the transmit mixer, Z102 on the exciter board, thus permitting generation of the transmitter signal.



CW SHAPING

For lower speed CW operation, it is desirable to have a slower ("softer") attack and decay time on the CW note, so the operator may switch C106 in parallel with C105, thus increasing the time constant. High speed CW is more easily copied when it has a fast ("hard") attack time, so C106 is then switched out of the circuit, at the operator's option.

Also, when the key line is high, CR104 is allowed to conduct, charging C102 and biasing Q106 on through R112 and R114. This lowers the base voltage of Q107, causing it to cease to conduct, and bringing the "R" line to ground potential. When the key line goes low, C102 discharges through R-112 and the R113-R114 and Q106 base-emitter junction parallel combination, delaying the turn off of Q106 and thus the return of the "R" line to a high value. This is to assure that there is a delay between the deenergizing of the "T" line and the reenergizing of the "R" line, so that certain switching functions can be accomplished without affecting either receive or transmit function.

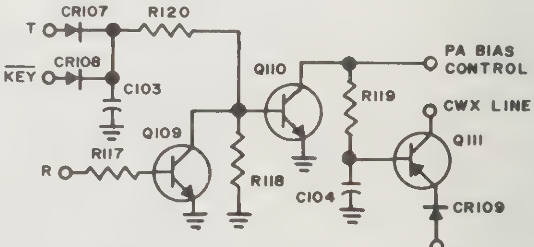


"R" LINE CONTROL

When semi-breakin is selected the CW timer holds the base of Q106 high for a brief time, preventing return to the receive condition between characters. When full breakin is selected, this feature is disabled.

When either the "T" line or the key line is high, voltage is applied to Q110, making it conduct. This brings the PA BIAS control line low, and turns on the bias to the power amplifier stages. Also, the base of Q111 goes low, allowing it to conduct. Q111 then conducts from the CW line through CR109, to the CWX line, energizing it.

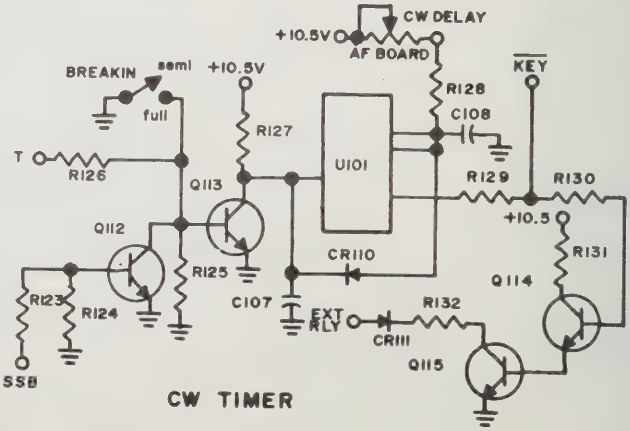
Should, however, the "R" line be high, voltage is applied through R117 to the base of Q109, causing it to saturate, and grounding the base of Q110, thus nullifying the effect of the "T" and key lines outlined above. Thus, there can be no output from the transmitter until the antenna switching circuitry has returned to the receive condition. This is an "interlock" function, which prevents damage to the receiver front end in the event of a control malfunction.



PA BIAS & CWX LINE CONTROL

CW TIMER

When the BREAKIN switch is in the SEMI position, energization of the "T" line applies a voltage to the base of Q113, causing it to saturate and the voltage at the input of timer U101 to go low. So long as this condition pertains, capacitor C108 is prevented from charging through R128 and the CW DELAY control by diode CR110. When the input is low, or until C108 is

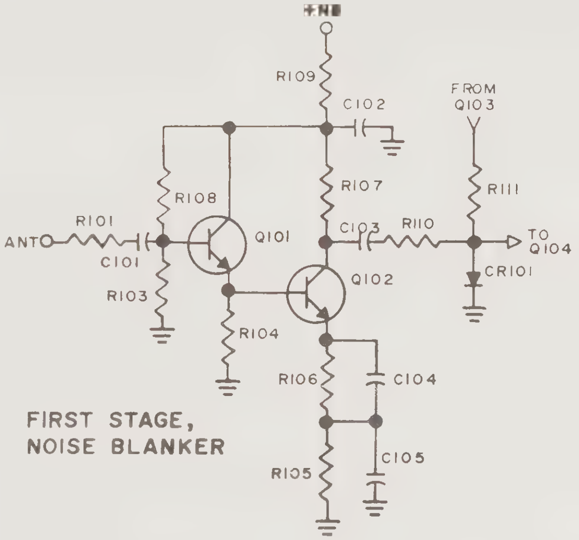


CW TIMER

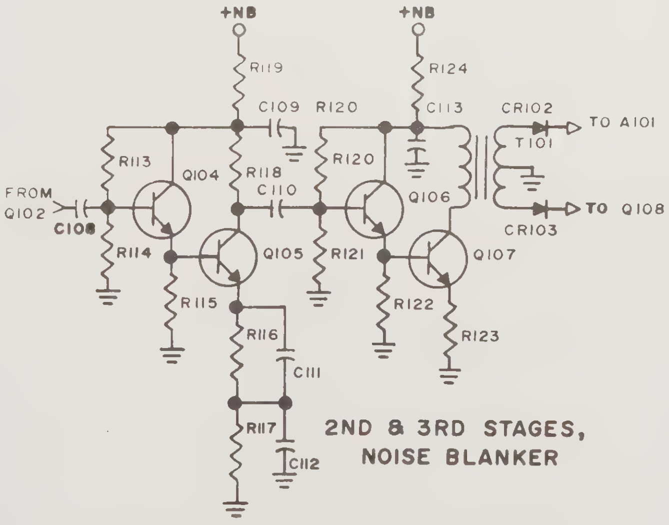
charged to 2/3 of the supply voltage after the input has returned high, the output of U101 is high. When this condition prevails, Q114 is turned on through R129 and R130. This in turn saturates Q115, which keys an external relay (such as for a linear amplifier, etc.) through protective diode CR111 and resistor R132.

Should FULL breakin be selected by the operator, the base of Q113 is grounded and the control input of the timer is never allowed to go low. Keying is then controlled by the key line, connected to the junction of R129 and R130. When the transmitter is keyed, this line goes high, and Q114 and Q115 are again saturated, grounding the external relay line. Note that few external amplifiers are capable of operating at full breakin, due to the slowness of the switching of the various relays, etc. Only those amplifiers especially built, or modified, for QSK operation should be used when full breakin is desired.

NOISE BLANKER



FIRST STAGE, NOISE BLANKER



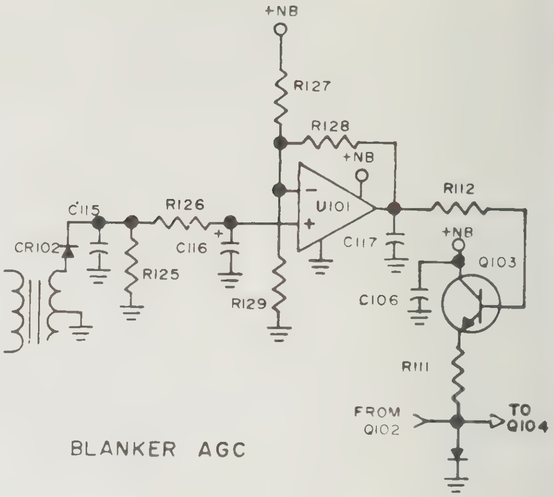
2ND & 3RD STAGES, NOISE BLANKER

The noise blanker signal input is taken directly from the receiver antenna input, and is coupled through isolation resistor-capacitor combination R101-C101 to the base of compound stage Q101 and Q102. Bias for that pair is derived from R108 and R103, with supply decoupling provided by R109 and C102.

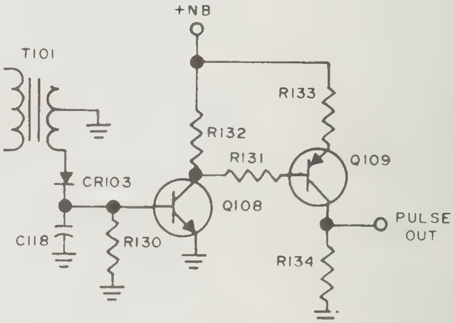
Frequency compensation is accomplished by R106-C104 and R105-C105 networks. Output is taken through C103, R110, and C108 to the base of the second compound stage. That stage, Q104 and Q105, operates in the same manner as the first one. In turn, this output is taken to yet a third compound stage, Q106 and Q107. This pair utilizes a conventional emitter network and untuned transformer T101 in its output. Note that the entire circuit is broadbanded--no selectivity is built in, so that noise pulses, no matter what frequency, will be accepted. The use of three compound stage amplifiers in cascade realizes a very high gain, for good sensitivity to the noise pulses.

The upper half of the secondary of T101 drives detector CR102, and after filtering by the C115-C116-R126 network, the resultant DC signal is applied to operational amplifier U101. The gain of this device is set by the ratio of R128 to R129 and its input threshold point is determined by voltage divider R127 and R-129 on the non-inverting input. The DC output of A101 is applied through current limiting resistor R112 to the base of Q103. Q103 supplies a current to PIN diode CR101 which serves as a variable attenuator, thus supplying AGC to the noise blanker receiver. The R125, R126, C115 and C116 network integrates all incoming signals to drive U101 and generate AGC to prevent strong signals (SSB, AM, or CW) from generating cross modulation in the blanking gate.

The bottom half of T101 drives detector CR103, which supplies a DC pulse to the base of Q108 when a noise pulse is received. When Q108 is thus turned on, it in turn causes PNP transistor Q109 to conduct, creating a blanking pulse across its load resistor, R134, which is coupled to the noise blanker gate in the I.F. circuit on the exciter board.



BLANKER AGC



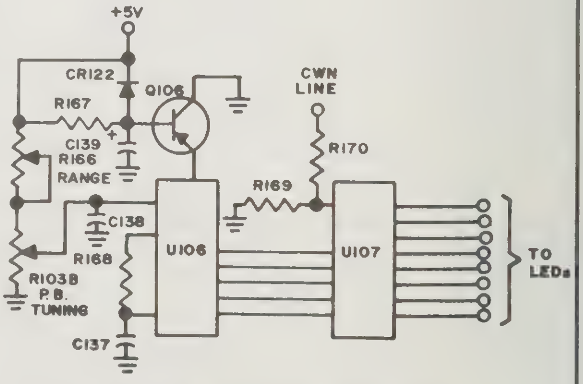
PULSE DETECTOR & PULSE AMPLIFIER

PASSBAND TUNING INDICATOR

A separate section of the front-panel PASSBAND tuning control, R103B, is used to develop a signal used to drive the series of LEDs that indicate the status of the tuning of the passband circuit. R166 is used in series with it to provide a range calibration adjustment, and the output, bypassed by C138, is coupled to the analog input of analog-to-digital converter IC U106, on the audio board.

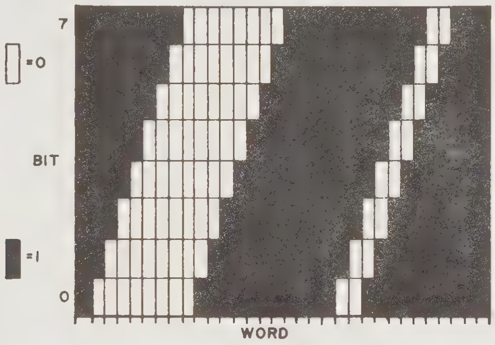
U106 is capable of free-standing operation, but to operate in such a manner, it must be provided with an R-C combination that sets the frequency of operation of the internal clock oscillator on the chip. R168 and C137 serve this function. The IC must also be properly initialized when it is first powered up. Q106 and its associated circuitry perform this task. C139 charges through R167 when power is first applied, thus allowing Q106 to conduct until nearly full charge is achieved, grounding the appropriate pins on the chip. Once C139 has charged, Q109 ceases to conduct, and the initialization pins on the IC are ungrounded, allowing it to operate. When power is removed, CR122 allows C139 to discharge rapidly.

The digitally-coded output of U107 is coupled to the inputs of PROM U107. The internally stored code in the PROM generates the required signals to drive the



PASSBAND TUNING INDICATOR

front-panel LEDs used as the passband tuning status indicator. When in the CW narrow mode, the CWN line energizes another input of the PROM, changing the code presented to the LEDs such that only two adjacent devices, representing the frequency range being passed by the PBT circuitry, are lighted.


UI07 MEMORY MAP

CUBIC
COMMUNICATIONS
SWAN DIVISION

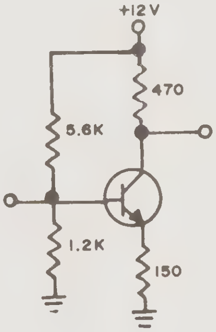
Servicing Hints

Transistor Troubleshooting........................7-1
Integrated Circuit Troubleshooting...............7-2
Measuring S+N/N Ratio............................7-2
Solid State Switching............................7-3
Two Tone Testing.................................7-3
Feedback Amplifiers..............................7-5
A Simple Impulse Noise Generator.................7-6
The Phase Locked Loop Synthesizer................7-7
Passband Tuning..................................7-8

TRANSISTOR TROUBLESHOOTING

Many techniques for troubleshooting transistor circuits have been evolved, of varying degrees of complexity and reliability. One system, perhaps the simplest and most reliable, is outlined here for your information. With it, you can determine, with only a few minutes thought and basic arithmetic calculations, what the normal operating voltages on any given transistor stage should be, and by comparing them with those actually measured, locate the fault in the circuit.

The key to the scheme is to remember that, for normal operation, the base-emitter junction of a transistor must be biased into conduction--and that when this is so, the voltage drop across that junction will be approximately 0.7 volts for silicon devices and about 0.25 volts for germanium.



To illustrate how the system works, lets take a simple amplifier stage (at the right). To start with, we know that the base current is very small, on the order of microamperes. For our purposes, then, we can ignore it, for it would be masked by circuit tolerances sufficiently that we couldn't measure a voltage drop caused by it with our usual methods. We therefore treat the bias network (the 5.6 kΩ and 1.2 kΩ resistors) as a simple voltage divider, and can compute the voltage that would appear at their junction (the base of the transistor). In this case, it works out to about 2.1 volts. Since we know that this silicon transistor has a base-emitter drop of about 0.7 volts, it follows that the eimtter voltage is that amount less than the base voltage, or about 1.4 volts. The emitter current can then be calculated by Ohm's law, and comes out to be about 9.3 mA.

We know that both the base current and the collector current flow through the emitter resistor, but the base current is so small that we can say for practical purposes that the emitter and collector currents are the same; therefore, the current through the 470Ω collector resistor is also about 9.3 mA, and the voltage drop across it is therefore about 4.4 volts. This means that the collector voltage would be 12 - 4.4, or 7.6 volts.

Actual measured values on a normally operating circuit would be close to these values, while a malfunctioning transistor would result in greatly different values being found. Remember, transistors do not get "weak", as do vacuum tubes (with the exception of some emitter-ballasted power devices) so when they fail, they do so catastrophically--they open or short, or both. Thus, the voltages found on their terminals will be greatly different than if it were operating normally. For example, a shorted base-emitter junction would result in no 0.7 volt drop between those terminals; an open base-emitter junction would leave the emitter at ground potential (no base current flowing, so no collector current, so no drop across the emitter resistor). And so on. It isn't hard to figure out just what is wrong with the device, although this isn't really necessary--if the voltages are not close to what you calculated, the device is probably bad.

INTEGRATED CIRCUIT TROUBLESHOOTING

Very little can be written concerning the finding of internal faults in integrated circuits. Often, the exact internal circuitry is both complex and unknown to the service technician, for manufacturer's data sheets generally show only an equivalent circuit at best, and at worst, only the pinout.

The data sheets often do, however, give a truth table or timing (waveform) diagram for digital circuits, or some other description, not of what the internal circuitry _is_, but rather what it _does_. This, then, is the approach we must use in troubleshooting them also.

In short, we troubleshoot IC's by the process of elimination; Are the voltages at each of the pins about what they should be? Is the proper signal going into it? Then, if the proper signal isn't coming out of it, it is bad and must be replaced. Never mind what is wrong inside--you can't get inside to fix it anyway, and time spent figuring it out is wasted. Just replace it.

A good voltmeter, a high impedance VTVM or better yet one of the newer digital voltmeters that will read to tenths of a volt or less, and a good oscilliscope are then necessary pieces of test equipment, and no shop should be without them. If you don't have them, you'll have to spend your time in speculation, and for the most part, the results will depend on guesswork. Either you'll use more time than necessary (and have to charge the customer for it) or you'll wind up missing the cause of intermittant problems, replacing perfectly good parts, or the like. In either case, the increased cost to the customer (or decreased profit to the shop) is something that you can hardly justify.

MEASURING SIGNAL TO NOISE RATIO

The procedure for measuring the signal to noise ratio of a receiver is really quite simple, and it is a check that should be made on every receiver before it leaves your shop--finding a weak receiver and fixing the problem while it's in the shop is one of the best ways to keep a customer happy, and it will pay dividends in repeat business.

Connect an unmodulated signal generator that has a 50Ω output impedance to the antenna terminals, and tune it for about 1 kHz beat note audible in the speaker. Adjust the generator's output for about 0.25 uV at the antenna terminals, so that the AGC circuit isn't activated. (Note that some generators are calibrated in terms of their _open circuit_ output voltage, not what is present when connected to a load--check yours out, so you can get the right setting.)

Connect an AC voltmeter across the speaker terminals, and adjust the volume control for a convenient reading on the meter. Note that reading. Now remove the signal generator connection and short out the antenna terminals without changing anything else on the receiver. Note that reading.

The first is the reading for the signal-plus-noise; the latter is the reading for noise only. The S/N ratio is 20 times the common log of the ratio between the first and second values. Some AC meters have log (or dB) scales and can be used directly.

SOLID STATE SWITCHING

One of the more useful innovations in recent years is the application of solid state switching to signals within receivers and transmitters. It allows switching to be accomplished at the site, without the need for routing long signal leads to a bulky, complex, and often troublesome multi-pole mechanical switch. Instead, a simple switch controls a DC voltage which in turn is applied to all of the points where switching of the signal is to take place.

The simplest form of a solid state switch is the diode. We know that when it is reverse biased, it presents a very high resistance--high enough to be considered an open circuit when used in low or medium impedance circuits. Modern manufacturing techniques have reduced the junction capacitance of switching diodes so low that they can also be used effectively in most high impedance circuits, too. Thus, a diode which has a reverse bias voltage greater than the peak value of any AC signal applied to it, so that it does not conduct during any part of the signal cycle, will act as an open circuit.

Conversely, modern diodes have a very low "on" resistance, so if forward biased with a current greater than the peak signal current, so that it conducts during all parts of the signal cycle, it passes the signal with little or no attenuation.

All that is needed, then, is a circuit arrangement of capacitors to block the DC but pass the signal, (and perhaps bypasses, if high-level RF is near) and resistors to apply the bias and switching voltages, and the diode itself.

Field Effect Transistors (FET's) can be used as even better solid state switches. Remembering that the source to drain path of the FET is a pure resistance, normally conducting, we can see that, as with any other resistance, current can flow in either direction (bilaterally), from source to drain or from drain to source, depending on the polarity of the applied voltage. Therefore, AC signals will pass.

Now, if we apply a voltage to the gate that is sufficient to "pinch off" the source-drain path, the effective resistance of that path increases greatly, effectively opening the circuit, and thence the signal path.

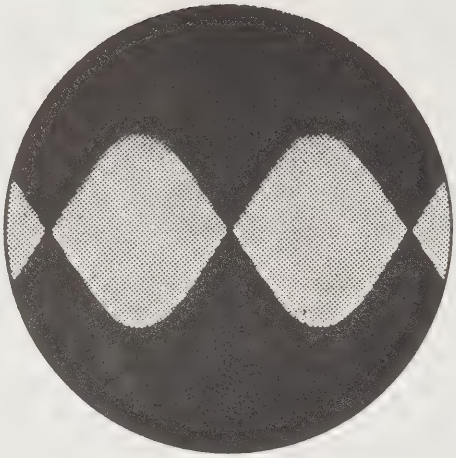
Normally, N-channel, depletion mode devices, with good bilateral characteristics in the source-drain path and low "on" resistance are selected for service in switching applications, although P-channel devices are also used.

All, then, that is needed for a FET switching circuit is the FET, suitable blocking capacitors and load and biasing resistors, just as in the case of the diode switch. In some instances, even these can be omitted, using the signal voltages themselves to provide any bias.

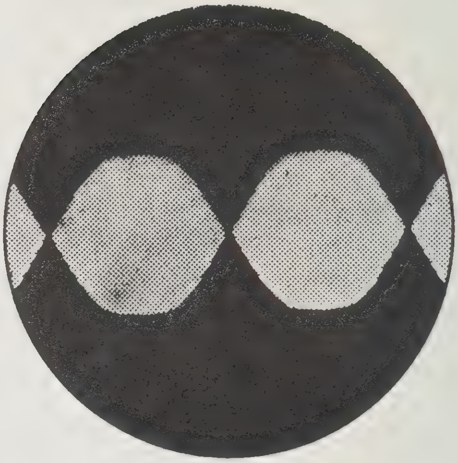
TWO TONE TESTING

Probably the most useful and yet least often performed test of a single sideband transmitter's performance a technician can make is the two tone test, and yet it is easily performed. The two tone test will disclose quickly and easily most of the major faults that are to be found in an operating transmitter by showing nonlinearities, improperly adjusted bias, overdriving, and many other problems.

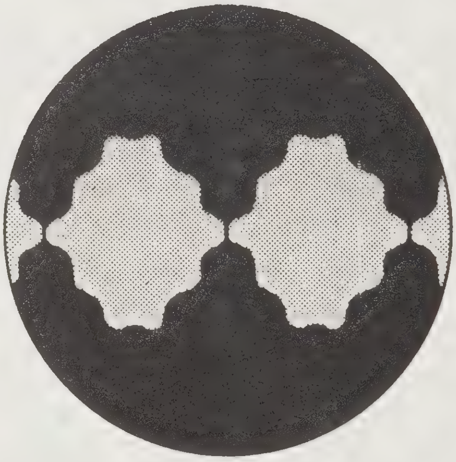
While a full discussion of the two tone test and its results is beyond the scope of this manual, a brief discussion is in order. More thorough



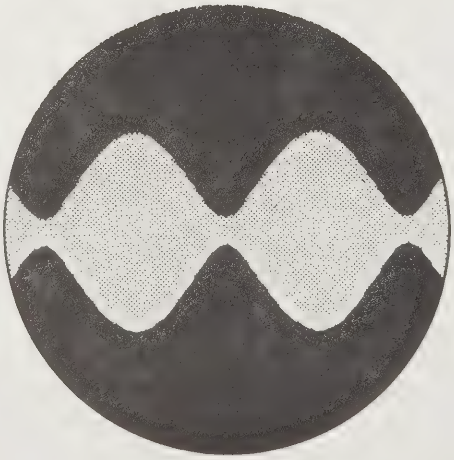
Two-Tone Pattern, Correctly Adjusted



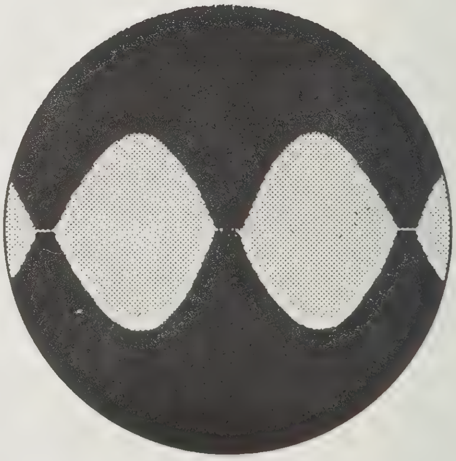
Two-Tone Pattern, Overdriven



Two-Tone Pattern, Carrier Present



Two-Tone Pattern, Unequal Tones



Two-Tone Pattern, Improper P.A. Bias

treatment is available in many competent references for those who wish to pursue the matter further.

Basically, this test is performed by applying two non-harmonically related audio tones, of equal amplitude and within the pass band of the radio, to the microphone input and observing the output across a dummy load on an oscilloscope. In this manner, the entire transmitter, from microphone jack to antenna terminal, is tested in one simple operation.

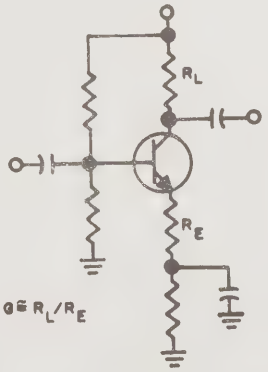
These tones can come from two audio signal generators, or better yet, from a two tone test set specifically designed for the purpose. Several such designs of adequate quality are included in amateur handbooks and other readily available literature. The tones must be very pure sine waves, and of equal amplitude, and not harmonically related to each other as mentioned.

The oscilloscope used to view the transmitter's output must be capable of displaying the frequencies at least as high as those in use, and preferably higher. A 30 MHz bandwidth instrument is adequate for all HF amateur bands, while the more common 15 MHz bandwidth instrument would serve for the 160, 80, 40 and 20 meter bands. Since any problems of nonlinearity, bias, etc., could be expected to show up on all bands, testing on one band only is usually considered adequate.

Several waveforms and their common causes are included here for reference, and others are available in current literature.

FEEDBACK AMPLIFIERS

The gain of individual transistors, within a given type number, unfortunately varies over a wide range. Often, this is satisfactory for the designer's purpose, but occasionally there arises a situation where the gain of the stage must be controlled more precisely. In that case, it is often advantageous to use an amplifier stage employing emitter degeneration (feedback) to set the gain of the stage. The accompanying diagram shows one such typical stage. The gain of the stage is approximately equal to the ratio of the collector load to the unbypassed portion of the emitter resistance. In the case of choke or tuned circuit loads, the value used in this calculation is the absolute value of the collector load impedance.

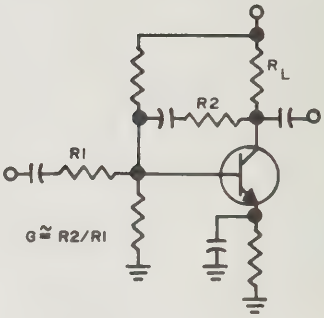


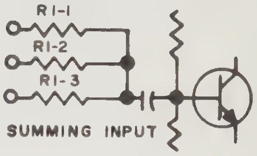
Utilizing this technique also raises the input and output impedances to higher values than would otherwise be the case, a condition that is often desirable. The bandwidth of the circuit can also be increased, under the right circumstances, although, under other conditions, it might be decreased instead. It is also possible to select the value of the capacitor bypassing a portion of the emitter resistance such as to provide frequency compensation for the inherent 6 dB per octave drop off in gain as the frequency is increased, that is a characteristic of all bipolar transistors.

Another type of feedback amplifier is shown in the second diagram. In this instance, the gain is approximately equal to R2 divided by R1, provided that the load resistance is much much greater than R2, and the impedance of the source feeding the circuit is much less than R1. The input impedance of this stage is equal to R1's value, and the output impedance is low. Since both input and output impedances can be quite low, the bandwidth may be very wide--100 MHz or greater.

Since the feedback path of the circuit (R2) allows just enough signal from the output to be fed back to the base to cancel out that coming in through R1 (and the feedback signal is out of phase with the incoming signal), little or no signal will be seen at the base. It is a "virtual" ground, so far as voltage is concerned!

This characteristic can be used to sum two or more signals, setting different gain values for each signal, if desired. The gain for each signal is proportional to the ratio of the feedback resistor (R2) to it's own input resistor only; the other input resistors do not affect the gain of that particular signal. It is interesting to note that this is a summing, not mixing, amplifier, often another useful property of the circuit.

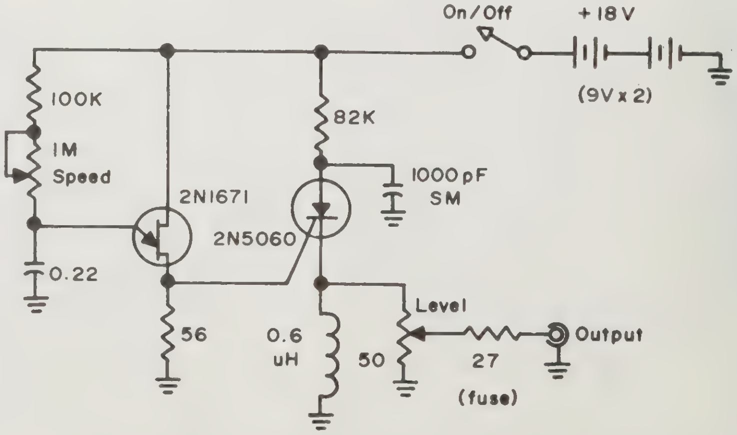




A SIMPLE IMPULSE NOISE GENERATOR

The noise blanker circuit used in SWAN radios, like all effective noise blankers, is designed to detect the presence of impulse noise above a certain threshold, and to mute, or blank, the output of the receiver for the duration of that pulse. (There is no circuit or scheme that can eliminate static or background hiss types of noise.) Since operation of the blanker circuit depends upon the presence of this high level impulse type of noise, it is difficult if not impossible to service the circuit unless a cource of such impulses is available. Efforts to service it using the commonly available test equipment lineup are ineffective, and may well lead to misadjustments--most often, this section of the receiver is never checked out, but merely accepted on faith.

It follows, then, that a simple, inexpensive instrument that generates this type of signal would be most useful in the service



Noise Generator

shop, yet commercial instruments are prohibitively expensive. For that reason, the circuit of such an instrument is included here, so that it may be built up locally and added to the shop equipment inventory for such service.

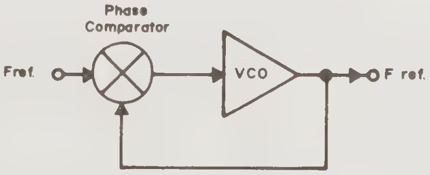
Referring to the accompanying diagram, we see that the 2N1671 is a unijunction transistor, connected in an oscillator circuit. The 0.22 mfd capacitor is charged through the 100K and 1M resistors, with the latter made adjustable to control the charging rate. When the charge reaches a certain level, the UJT fires, discharging the capacitor, and generating a sharp pulse of current through the 56 ohm resistor. This in turn fires the 2N5060 delivering thereby a relatively large current pulse through the inductor. This pulse is used to simulate the impulse noise. Level is adjusted by the 50 ohm potentiometer, and the small wattage 27 ohm series resistor acts as an output fuse, for protection from the circuit to which it is connected. Power is shown as two 9-volt batteries in series, although a suitable AC supply delivering approximately 18 volts DC at a few milliamperes would also be suitable.

THE PHASE LOCKED LOOP SYNTHESIZER

Electronic phase locked loops (PLL) have been in use since the 1930's where they were used for radar synchronization. In later years, they came into more widespread use as the synchronization system in television receivers and other applications. It is only in recent years that they have found widespread use as frequency synthesizers, where they have found many applications in satellite communications systems, airborne navigation systems, FM communications, and, most recently, in HF SSB communications equipment.

In the basic PLL system, the output of a local voltage controlled oscillator (VCO) is compared with an incoming, or reference, signal, and if the two are not of the same frequency and phase, an error signal is generated to tune the VCO in such a direction, and to the required degree, that they become so. This is the system used in television sync systems, and is shown in the accompanying diagram.

The usefulness of such a simple system is limited, however, to such synchronization tasks; however, it was found that, if a means of dividing the output frequency of the VCO could be inserted in the path between the VCO and the comparator, the output of the VCO could be made to automatically assume a frequency that was a multiple of the reference frequency, that multiple being equal to the division factor of the divider circuit.

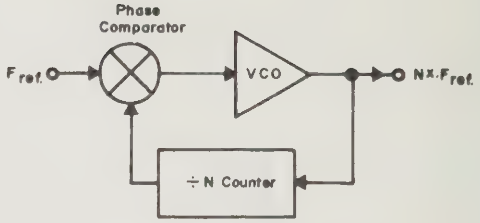


BASIC PHASE LOCKED LOOP

By making the divider circuit programmable, that is, making it so that the division ratio could be varied at will, and making the VCO of an appropriate frequency range, true frequency synthesis could be achieved; that is, the frequency of the VCO could be made to vary, in integer multiple steps of the reference frequency. This scheme is shown on the following page.

Digital circuitry lent itself quite handily to the implementation of the frequency divider, in that a number of available chips could be used as programmable dividers, or in the present terminology, "divide-by-N" counters, where N is the dividing integer. VCO's of various configurations were easily achieved, using varicaps as the tuning element.

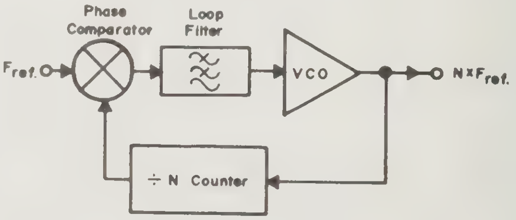
Initial PLL designs used phase comparators constructed out of discrete components, and were often very complex and difficult to "tame".



BASIC MULTIPLIER PLL

The development of specially designed integrated circuit phase comparators that were basically digital mixers greatly alleviated many problems, but created another. In such a mixer, leak-through of the reference frequency components to the error, or tuning line to the VCO is common unless specific steps are taken to prevent it. Such leak-through results in frequency modulation of the VCO at the reference frequency, and if held low enough, will be unnoticable in the reception of AM or FM signals, but is quite difficult to reduce to a level where SSB or CW signals can be received without objectionable distortion caused by this FMing. The means taken to reduce this distortion is the includion in the loop, between the phase comparator and the VCO, of a filtering arrangement to reduce the amplitude of the reference frequency feedthrough to the VCO. Usually a low pass filter is used for this purpose.

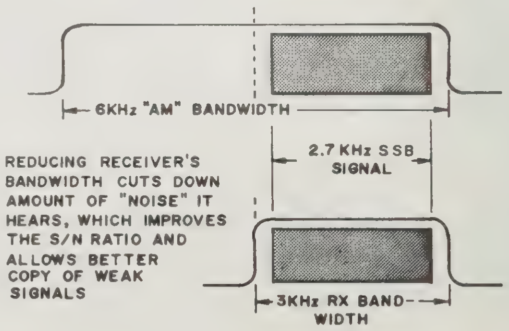
Since the loop filter contains considerable capacity to ground, of necessity, and its cutoff frequency is below the reference frequency, it controls the time required for the loop to reachieve lock after a change has been introduced, such as when the divide ratio of the counter has been changed. At low reference frequencies, lockup time can become objectionably long. In the Astro 102BX, a synthesizer is used to generate the different band frequencies which are mixed with the PTO output to produce the local oscillator signal. The reference frequency is sufficiently high that a relatively simple loop filter is all that is required.
Setting the division ratio into the programmable counter is accomplished by the BAND switch, in conjunction with a diode matrix and programmable read-only memory integrated circuit.



PRACTICAL MULTIPLIER PLL

PASSBAND TUNING

It is a fact of communications life that the narrower the bandwidth of a re-



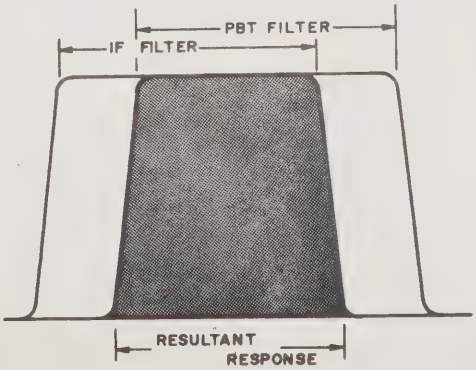
REDUCING RECEIVER'S BANDWIDTH CUTS DOWN AMOUNT OF "NOISE" IT HEARS, WHICH IMPROVES THE S/N RATIO AND ALLOWS BETTER COPY OF WEAK SIGNALS ceiving system, the more readily a signal of a given strength can be received and understood. This phenomenum occurs because what we really hear is the signal to noise ratio, rather than the absolute signal strength, and the narrower the received bandwidth, the less noise is received; hence a better signal to noise ratio and a more understandable recovered signal. The limiting factor, of course, is the bandwidth required for intelligibility of the desired signal. This is one of the chief advantages of single sideband over other communications methods. Since the necessary bandwidth is less than half that necessary for AM, for instance, it will result in more than twice the signal to noise ratio than an AM signal of comparable power, due solely to the reduction in received noise.

Studies show that the minimum bandwidth required for intelligible transmission of the human voice is about 2.5 kHz, although one can "get by" on a bandwidth of as low as 2.0 kHz, if a bit more degredation of the voice quality can be tolerated. In actual fact, however, these are average figures, and individual voice requirements vary widely. It has been the practice (and still is, in many cases), to use a fixed bandwidth, set by a high quality crystal filter in the IF of the receiver, to improve the signal to noise ratio. The response of a typical filter is shown to the right.

TYPICAL SSB FILTER CURVE

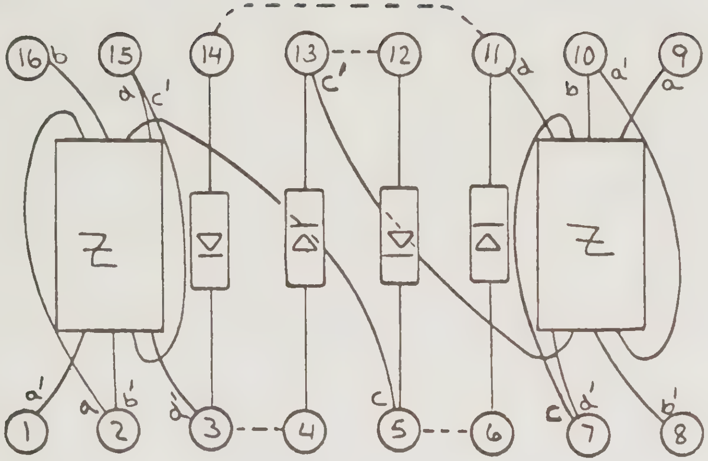
Since individual voices vary in their requirement for bandwidth, however, it would be advantageous in many instances, to be able to vary the bandwidth to lower limits. By being able to cut off the top or the bottom of the passband, thus reducing the bandwidth, it is possible to reduce the effects of close-by interfering signals, often at acceptable levels of degredation of the desired signal quality. Crystal filters, however, do not lend themselves to such operator adjustment.

Quite recently, a scheme for accomplishing this has been developed; it is called "Passband Tuning". In this scheme, the basic bandwidth is set by a high-quality crystal filter in the IF as before. But, after that filter, a circuit is included to translate that IF frequency to another frequency, usually higher. The signal is then passed through another crystal filter, of similar bandwidth as the first, and then translated back to the first IF frequency. If the frequency of the oscillator used to translate the signal to and back from the second IF is such that the passband of the two filters line up, no change in the bandwidth is noticed; however, if this oscillator frequency is made variable, the passbands of the two filters no longer lines up, and either the upper or the lower end of the passband of the first filter is cut off by the second, as shown in the accompanying diagram. Since the frequency of this passband tun-

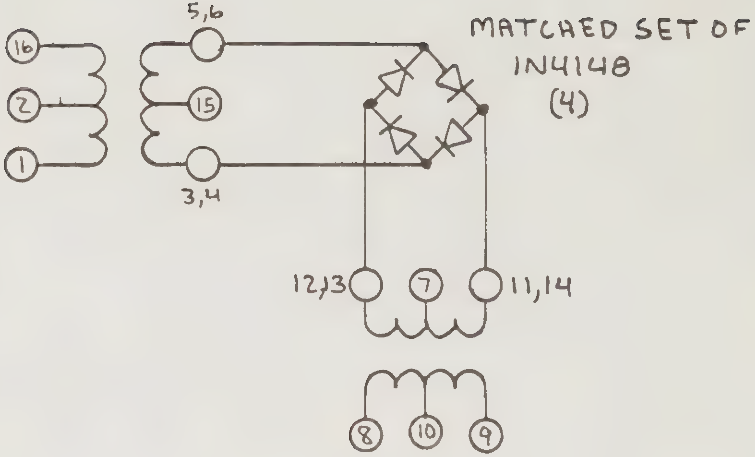


ing oscillator is continuously variable, we have achieved continuously variable bandwidth. Skillful operation of this system by the operator can result in useful copy of signals that would otherwise be uncopiable due to heavy interference from nearby signals. Such a scheme, with a unique visual indication of the bandwidth to which it is adjusted, is included in the Astro 102BX.

The following two pages provide data on the diode ring double balanced mixers used for Z2101, Z2102, and Z5101. The first type uses discrete diodes and the second uses a monolithic diode array. The two units are interchangeable.



TOP VIEW



MATCHED SET OF
1N4148
(4)

5. THE FOLLOWING CONNECTIONS MADE EXTERNALLY
3 TO 4, 5 TO 6, 12 TO 13, 11 TO 14
4. MOUNTED ON 16 PIN SOCKET P.N. CH6170-16
3. WIRE COLORS
a - RED
b - BLUE
c - GREEN
d - CLEAR
2. USE (2) Z BEAD TRANSFORMERS, TYPE 1
1. USE (4) MATCHED DIODES 1N4148
NOTES

SWAN ELECTRONICS

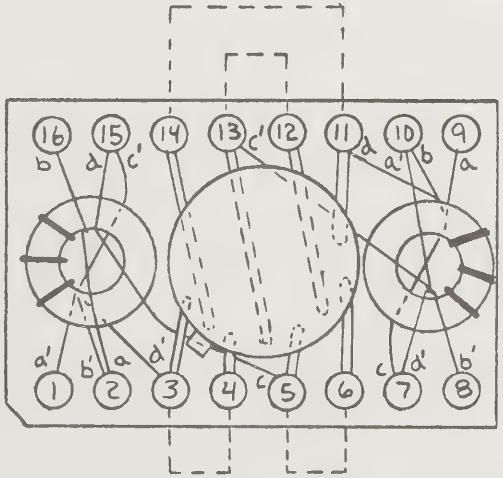
SCALE: NONE
DATE: 5-26-77
APPROVED BY: Jcw
DRAWN BY C.DAVIDSON
REVISED

MIXER

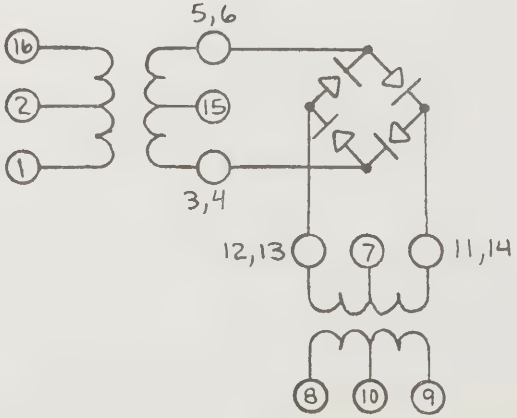
DRAWING NUMBER
0800-2004

REVISIONS

LTR	DESCRIPTION	DATE	APP'D
A	REVISED DWG. TO SHOW CA3039 DIODE PACKAGE INSTEAD OF SEPARATE DIODES CD 3-5-80	3-5-80	



TOP VIEW



4. (1) CA3039 MODIFIED PER DWG. NO. 0800-5404
3. (2) RF TRANSFORMERS 014-063
2. WIRE COLORS
   a - RED
   b - BLUE
   c - GREEN
   d - CLEAR
1. THE FOLLOWING CONNECTIONS MADE EXTERNALLY
   3 TO 4, 5 TO 6, 12 TO 13, 11 TO 14

NOTES

CUBIC COMMUNICATIONS, INC.
905 AIRPORT ROAD • OCEANSIDE, CA. 92054

SCALE: NONE	APPROVED BY:	DRAWN BY C. DAVIDSON
DATE: 3-4-80		REVISED

DOUBLE BALANCED MIXER

DRAWING NUMBER
0800-2004A

Servicing Data

Test Equipment Required....8-1
Initial Current and Light Checks....8-2
PTO Mode Switching and RIT Operation Check....8-2
PTO Range and Display Operation Check....8-3
Display Frequency Accuracy Check....8-3
Transmitter Operation Check....8-3
ALC Operation Check....8-4
Modulation Operation Check....8-4
Receiver Operation Check....8-5
Noise Blanker Operation Check....8-6
Passband Tuning Operation Check....8-6
CWN Operation Check....8-7
PTO Operation Check....8-7
External Speaker Check....8-7
External L.O. Check....8-7

Alignment Procedure
  Initial Setup....8-8
  Voltage Regulator....8-8
  Band VCO....8-8
  Band Reference Oscillator....8-8
  Carrier Oscillator....8-9
  Passband Tuning Oscillator....8-9
  Counter Reference Oscillator....8-9
  Passband Tuning Indicator....8-9
  PTO and RIT....8-9
  Exciter I.F....8-10
  AGC and S Meter....8-10
  CWN Filter....8-10
  Notch Filter....8-10
  ALC and Meter....8-11
  Preselector and L.O. Bandpass Filters....8-12

Level Diagrams
  L.O. Board....8-14
  Exciter Board....8-14
  Distribution Board....8-15
  Audio Board....8-15
  PTO Control Board....8-16
  Noise Blanker Board....8-16
  Counter Board....8-16
  Preselector Board....8-17

Band Synthesizer Memory Map....8-17
PBT Indicator Memory Map....8-17

Layouts/Schematics
  PTO Control Board....8-18, 8-19
  Preselector Board....8-20, 8-21
  Low Pass Filter/Band Selector Board....8-22, 8-23

L.O. Board (Layout)....................................8-24
Counter/Display Boards..................8-25, 8-26, 8-27
Audio Board..................................8-28, 8-29
Distribution Board...........................8-30, 8-31
Front Panel Right Board......................8-32, 8-33
Passband Indicator Board.....................8-34, 8-35
Power Amplifier Board........................8-36, 8-37
SWR/Wattmeter Board..........................8-38, 8-39
Noise Blanker Board..........................8-40, 8-41
PTO Board....................................8-42, 8-43
Exciter Board................................8-44, 8-45
L.O. Board...8-46

Main Chassis/Interconnect Schematic..................8-47

TEST EQUIPMENT REQUIRED

A certain amount of quality test equipment is necessary to service today's sophisticated communications equipment. Below is a list of that required for the ASTRO 102BX.

-Microphone, high impedance

-Wattmeter, 0-200W scale

-Dummy load, non-reactive, 50 ohms, capable of >200W dissipation

-Light bulb load, 100W, 10 ohms cold

-Oscilloscope, 30 MHz minimum bandwidth, with x10 attenuator probe to be used for all measurements

-Frequency counter, at least 100 mV sensitivity, high impedance input, counts to 44 MHz or higher

-VTVM or DVM with high input impedance to be used for all AC and DC measurements

-Two-tone generator (See the "Servicing Hints" section of this manual)

-Noise generator (See the "Servicing Hints" section of this manual)

-Signal generator, 1.5-30 MHz, calibrated output

-Audio generator, 300-3000 Hz

-Power supply, 11-14V DC adjustable, 22 ampere or greater regulated output, with current metering

-"T" coaxial connector

-Attenuator for monitoring transmitter output with counter and oscilliscope

INITIAL INSPECTION AND SET-UP

Inspect the unit for physical damage; i.e., broken components, knobs, switches, meter, panel, heatsink, controls, circuit boards, cabinet, etc.

burned wires, components and boards; loose and improperly mounted plugs; frayed or pinched wires; loose hardware, etc.

Connect a suitable power supply to the POWER 13.8 volt connector on the rear apron of the unit. Connect the wattmeter, dummy load and "T" connector with the attenuator to the ANTENNA connector. Connect a suitable microphone to the KEY/MIC jack on the front panel.

Set the front panel controls as follows: AF GAIN off; RF GAIN, fully clockwise; IF GAIN fully clockwise; AGC DELAY fully counterclockwise: PASSBAND 12 o'clock; NOTCH FILT fully counterclockwise; MIC GAIN fully counterclockwise; PTO MODE to PTO A; MODE to LSB and BAND to 20.

INITIAL CURRENT AND LIGHT CHECKS

With the power off, the total current drain should be less than 15 ma. Turn the AF GAIN control to turn the power on, and the total current drain should be less than 2 amperes, the meter light should come on, the display should light, and the PTO A and PASSBAND KHZ lights should come on.

Key the unit with the microphone button. The total current drain should increase between 1 to 1.5 amps.

PTO MODE SWITCHING AND RIT OPERATION

Set the PTO MODE switch to EXT, and the display should read 11000.1 ± 1. Set the MODE to USB, and the display should read 10996.8 ± 1. Both PTO lights should be off.

Set the PTO MODE to A and adjust PTO A for 14260.0 with the RIT off (RIT light off). The PTO A light should be on and the PTO B light off. Key the unit. The display frequency should not change, and there should be no change in the PTO lights. Unkey the unit.

Set the PTO MODE switch to B and adjust PTO B for 14280.0 with the RIT off. The PTO B light should be on and the PTO A light should be off. Key the unit, and the display frequency should not change, with no change in the PTO lights. Unkey the unit.

Set the PTO MODE switch to REC A-XMIT B. The PTO A light should be on, the PTO B light should be off, and the display frequency should be 14260.0. Key the unit; the PTO A light should go off, the PTO B light should be on, and the display frequency should change to 14280.0. Unkey the unit.

Set the PTO MODE switch to REC B-XMIT A; the PTO A light should be off, the PTO B light on. The display frequency should be 14280.0. Key the unit, and the PTO A light should come on, and the PTO B light extinguish. The display frequency should be 14260.0 Unkey the unit.

Set the PTO MODE switch to PTO A, MODE to LSB. The display frequency should change less than 0.2 kHz. Set the RIT knob pointer to 0 and push it in to turn the RIT on. The RIT light should come on, and the display frequency should change less than 0.2 kHz. Adjust the RIT control from full counterclockwise to fully clockwise. The display frequency should vary between ± 0.8 to 1.5 kHz. Key the unit. The RIT light should go off, and the display frequency should be the same as if the RIT were off, and the RIT control should have no effect on the display frequency. Unkey the unit. Push the RIT control to turn it off. The RIT control should have no effect on the display frequency. Set the MODE switch to USB.

Set the PTO MODE switch to B and make the same checks as for PTO A.

PTO RANGE AND DISPLAY OPERATION

Set the PTO MODE switch to PTO A, turn the PTO A knob counter clockwise to its limit. The display frequency should be between 13940.0 and 13880.0. Turn the PTO A knob clockwise to its stop, and the display frequency should be between 14560.0 and 14620.0. Set the BAND switch to 10D. The display frequency should be stable. Set the BAND switch to 20.

Set the PTO MODE switch to PTO B and repeat the range checks for PTO B.

Set the BAND switch to 10B, and adjust PTO B for a display frequency of 28888.8 and check that the segments and decimal point are properly lighted on the display.

DISPLAY FREQUENCY ACCURACY

Connect a frequency counter to the "T" coupling and attenuator in the antenna line. Disconnect the microphone, and connect an audio generator to the EXT MOD jack on the rear apron of the unit. Connect an SPST switch to the KEY jack on the rear apron.

Adjust the audio generator's output to 1000 Hz ± 10 Hz. Key the unit with the SPST switch and adjust the MIC GAIN control and audio generator output level for approximately 20 watts output, or until the frequency counter has a stable count. Check the following:

That the carrier (output) frequency equals the display frequency minus 1.0 kHz in the LSB mode, ± 0.4 kHz.

That the carrier (output) frequency equals the display frequency plus 1.0 kHz in the USB mode, ± 0.4 kHz.

That the carrier (output) frequency equals the display frequency plus 0.8 kHz in the CW and CWN modes, ± 0.4 kHz.

TRANSMITTER OPERATION

Disconnect the frequency counter from the "T" connector in the antenna line and connect the oscilloscope in its place. Keep the attenuator in the line to protect the oscilloscope input circuitry. Disconnect the audio generator from the EXT MOD jack and connect the two-tone generator to the EXT MOD jack.

Check the power output and the total drain current on all bands in the CW mode, with the MIC GAIN control fully clockwise, and with the output frequency adjusted to the high end of each band. The power output should be 100 to 120 watts on all bands and the total current drain from the power supply should be 18 amps or less at 100 watts output. Power output should vary from less than 5 watts to full power as the MIC GAIN control is varied from fully counterclockwise to fully clockwise. There should be no output until the unit is keyed with the SPST switch.

Check the meter accuracy in the FWD position at 100 watts output on the 20 meter band. Accuracy should be ± 10%. Disconnect the coax at the ANTENNA connector on the rear apron, and set the METER switch to REFL, the MIC GAIN fully clockwise, and key the unit momentarily. The panel meter should indicate in the red area. Reconnect the coax to the ANTENNA connector. Set the meter switch to ALC, key the unit, and vary the MIC GAIN control from fully counterclockwise to fully clockwise. The meter reading should vary from approximately 0 to +20 or higher. Unkey the unit, and the meter should read approximately 0 in any of the METER switch positions.

ALC OPERATION

Disconnect the 50 ohm dummy load and substitute a standard 100W light bulb (10 ohms cold) as the load. Check the power output with the MIC GAIN control fully clockwise on all bands. Power output should be 40-70 watts if the ALC circuit is operating properly.

With the antenna connector open (i.e., no load), the total set current in transmit should not exceed 12 amperes. If so, the ALC circuit is not working properly.

NOTE: The purpose of the ALC is to limit power output and therefore PA device dissipation to safe limits, by limiting collector current. If there is too much ALC control, it will not be possible to obtain full output into loads with SWRs less than 1.5:1. If there is too little ALC control, then the PA transistors can draw excessive current, especially when the load is poorly matched, and possibly go into oscillation, causing them to overheat and fail. The temptation to increase the power output of the transmitter by reducing ALC control must be resisted--to do so will almost guarantee the failure of expensive components in the PA.

MODULATION OPERATION

Check the two-tone signal on each band and in both USB and LSB modes for crossover distortion, carrier leakage, and spurious outputs. None of the above should be detectable in the output (see the "Servicing Hints" section of this manual for examples and procedures for making the two-tone test.) Output should go to 100 W. PEP without distortion being introduced into the two-tone pattern.

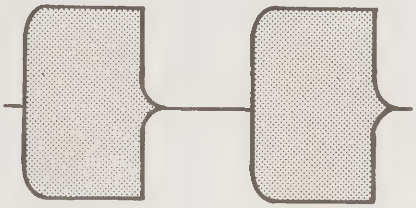
NOTE: If a two-tone source is not available, a close examination of a modulated output display using a high quality microphone and/or monitoring the signal on another radio should show up any major discrepancy, although not so accurately as does the two-tone test.

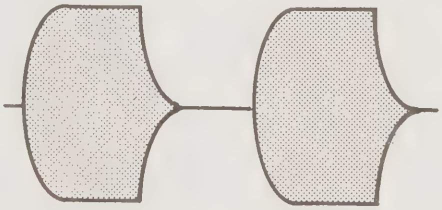
Two-tone and microphone modulation checks should be made with the speech processor off. Disconnect the two-tone generator and reconnect the microphone. The MIC LEVEL trimpot (R138) on the audio board should be adjusted for 100 W. PEP output with the MIC GAIN control set at 10 o'clock and speaking into the microphone that is to be used with the unit, in a normal voice. Note the average power reading on the wattmeter. Switch the SPEECH PROCESSOR on and note that the average power reading increases by approximately 20 watts. It should also be noticable on the oscilloscope display that the peak to average modulation levels are less with the speech processor on than when it is off.

With the VOX-PTT switch in the VOX position, adjust the VOX GAIN trimpot (R140) on the audio board so that speaking into the microphone instantly keys the unit. Adjust the ANTI TRIP trimpot (R149) on the audio board so that audio from the speaker does not key the unit. Varying the VOX DELAY trimpot (R144) on the audio board should vary the unkeying delay time from 0 to 5 seconds. There should be no loud popping noises in the speaker as the unit keys or unkeys. Disconnect the microphone.

Set the MODE switch to CW, the BRK IN switch to FULL, the CW HARD-SOFT switch to HARD. Key and unkey the unit rapidly several times; the receiver audio should return the instant the unit is unkeyed. There should be no objectionable pops or clicks from the speaker when keying or unkeying at any power output setting. Set the BRK IN switch to SEMI. Key and unkey the unit several times. There should be a 0.5 to 6 second delay between unkeying and receiver turn on that varies with the setting of the CW DELAY trimpot (R174) on the audio board.

Set the BRK IN switch to FULL, key the unit, and adjust the MIC GAIN control for a reading of 75 on the meter with the METER switch in the ALC position. Rapidly key and unkey the unit and compare the oscilloscope display to the illustration presented here, to check for proper waveshaping. Set the CW HARD/SOFT switch to SOFT and make the same check for waveshape.


"Hard" keying


"Soft" keying

The SIDETONE trimpot (R124) on the audio board should vary the level of the audio tone heard in the speaker when the transmitter is keyed. The tone should be present only in the CW or CWN MODE of operation, and should not be heard with the unit unkeyed. Output should vary from no output to approximately 1 watt of audio with the setting of the sidetone level control.

Connect one lead of a 12 volt lamp (less than 200 ma) to +12 volts from the power supply and connect the other lead from the lamp to the center contact on the EXT RLY jack on the rear apron of the unit. The lamp should light only when the unit is in transmit in any mode.

RECEIVER OPERATION

Disconnect the microphone, the SPST switch, wattmeter, "T" connector and attenuator, and the dummy load from the unit. Connect an RF signal generator to the ANTENNA jack. Connect an AC voltmeter and frequency counter across the speaker terminals. Set the MODE switch to USB, VOX/PTT switch to PTT, MIC GAIN control fully counterclockwise, and PTO MODE to PTO A.

Check for 0.35 uV sensitivity at 10 dB S+N/N ratio on all bands. See the "Servicing Hints" section of this manual for the procedure. Check the sensitivity with the PTO MODE switch in the PTO B position also. Check that the sensitivity is better than 1.0 uV at 10 dB S+N/N ratio with the input voltage set at 11V DC on all bands.

Set the RF generator output to 1 uV and note the S meter reading. Tune the unit to each internally generated spurious signal and note that they each should be less than the S meter reading noted above.

Check that the S meter reads S 9 ± 1 S unit for 50 uV input signal at a supply voltage of 13.8 V DC. The S meter reading should vary with the input signal. Changing the input signal from 1 uV to 0.1 V should result in 6 dB or less increase in audio output (6 dB is a voltage ratio of 2). The

AF AIN control should vary the audio output from inaudible to 3.5V RMS before clipping occurs, with 0.5 uV of input signal.

Turning the IF GAIN control fully counterclockwise should increase the S meter reading to full scale and decrease the units sensitivity. Set the IF GAIN control back to its fully clockwise setting.

Turning the RF GAIN control counterclockwise should decrease the units input sensitivity and the S meter reading should decrease. With the RF GAIN fully counterclockwise, input sensitivity should be reduced approximately 50 dB. Set the RF GAIN control back to its fully clockwise position.

With the AGC DECAY control fully counterclockwise, the unit should have no AGC delay time. This is indicated by the S meter reading increasing and decreasing instantly as the PTO is tuned across a signal. As the AGC DECAY control is advanced clockwise, the amount of time required for the AGC to decay should increase to approximately 15 seconds at fully clockwise. This is indicated by the slow drop in S meter readings after tuning off of a signal. Set the AGC DECAY control back to its fully counterclockwise position.

Adjust the RF generator so that the speaker audio tone is 1650 Hz ± 20 Hz. Adjust the NOTCH FILTER control for minimum speaker audio. The audio signal should decrease approximately 25 dB. Adjust the RF generator so that the speaker audio tone is 2500 Hz ± 20 Hz. Adjust the NOTCH FILT control for minimum speaker audio. The audio signal should decrease about 25 dB. Repeat the foregoing at 400 Hz ± 10 Hz audio tone. The audio signal should decrease by about 25 dB. Set the NOTCH FILT control back to its fully counterclockwise position.

NOISE BLANKER OPERATION

To check the noise blanker circuit, connect a pulse type of noise generator to the antenna connector along with the signal generator using the "T" connector. If a noise generator is not available, see the "Servicing Hints" section of this manual for the schematic diagram of a simple inexpensive set that may be constructed and kept for future use.

Tune in a 0.5 uV signal from the generator; turn on the noise generator and set the noise level to S 3 to S8 with the NOISE BLANKER switch off. Turn the NOISE BLANKER switch on; the signal should now be audible above the residual noise. Check on all bands.

Tune the signal generator to another frequency at least 1 MHz away from the set frequency and modulate the signal generator at 100% modulation. Increase the generator output to about 100,000 uV and at this point the noise level should come back up to S4 to S8; switching the NOISE BLANKER switch from on to off should have no effect.

PASSBAND TUNING OPERATION

With the PASSBAND control set at 12 o'clock, all eight PASSBAND KHZ lights should be on in LSB, USB, and CW modes. In CWN mode, only the 1.5 and 1.8 lights should be on. In LSB, USB and CW mode, as the PASSBAND control is turned clockwise, the PASSBAND KHZ lights should go off sequentially proceeding from left to right. At about the 4 o'clock position of the control, all lights should be out. As the PASSBAND control is turned counterclockwise from the 12 o'clock position, the PASSBAND KHZ lights should go off sequentially proceeding from right to left. At about the 8 o'clock position of the PASSBAND control, all lights should be out. Set the PASSBAND control back to the 12 o'clock position.

CWN TEST

Set the MODE switch to CW, and adjust the RF generator for 50 uV output and adjust the frequency for a 1650 ± 20 Hz speaker audio tone. Switch the MODE switch to CWN and note that the S meter reading changes by one S unit or less. Set the MODE switch to CW.

Adjust the RF generator frequency for a 2700 Hz ± 20 Hz speaker tone and set the RF generator output for S9 on the S meter. Turn the PASSBAND control counterclockwise until the S meter reading drops to less than S5. At this point, only the 2.7 and/or 2.4 PASSBAND KHZ lights should be off. Adjust the RF generator frequency for 400 Hz ± 20 Hz tone in the speaker and set the generator output for an S9 reading on the S meter. Turn the PASSBAND control clockwise until the S meter reading drops to less than S5. At this point only the 0.6 and/or 0.9 PASSBAND KHZ lights should be off. Set the PASSBAND control back to its 12 o'clock position.

PTO CHECK

Set the PTO MODE switch to PTO A and adjust the RF generator for 50 uV output. Tune PTO A across the signal and listen to the speaker audio tone for any indication of tuning backlash. Tune PTO A for a 1000 Hz ± 20 Hz tone in the speaker.

Set the PTO MODE switch to PTO B. Tune PTO B across the signal and listen to the speaker audio tone for any indication of tuning backlash. Tune PTO B for a 1000 Hz ± 20 Hz speaker tone. Both PTOs are now set to the same frequency and there should be no interference between the two. Slowly tune PTO A back and forth while listening for a change in the audio tone. There should be no detectable change.

EXTERNAL SPEAKER CHECK

Connect an external speaker to the EXT SPKR/PHONE jack on the rear apron of the set. The internal speaker should mute and the added speaker now should have audio output.

EXTERNAL L.O. CHECK

Disconnect the RF generator from the ANTENNA connector and reconnect it to the EXT LO jack on the rear apron of the unit. Adjust the RF generator frequency to 30 MHz and approximately 0 dBm output. Set the MODE switch to LSB, PTO MODE switch to EXT, and BAND to 15. The display frequency should be the RF generator output frequency minus 9000.0 kHz ± 0.1 kHz. The display should be stable and tuning PTO A or PTO B should have no effect on the display frequency.

This completes the operational test of the unit. Disconnect all test equipment and reinstall any covers that may have been removed.

ALIGNMENT PROCEDURE

Alignment should be performed only after making a complete performance check of the unit. It is advisable to then align only those areas of the unit necessary to bring performance up to specification. Also, the replacement of a defective component will usually restore the units performance without requiring any realignment, in most cases. Following this procedure will save time and help to avoid the possibility of inadverdantly misaligning a section of the unit.

INITIAL SETUP

Remove the top and bottom covers of the unit. Set the front panel controls as ollows: AF GAIN to off; RF GAIN to fully clockwise; IF GAIN to fully clockwise; AGC DECAY to 12 o'clock; PASSBAND to 12 o'clock; NOTCH FILT to fully counterclockwise; MIC GAIN to fully counterclockwise; PTO MODE to PTO A; MODE to LSB; BAND to 160 and PTT/VOX switch to PTT.

Connect a suitable power supply to the POWER 13.8 V connector on the rear apron of the unit.

VOLTAGE REGULATOR ADJUSTMENT

Turn the AF GAIN control clockwise to turn the power on. Adjust the power supply voltage for 14 V DC.

Measure the voltage at the collector of Q116 on the distribution board and adjust the 10.5 V SET trimpot (R141) on the same board for a reading of 10.55 ± 0.05 V DC on the voltmeter. Vary the power supply voltage from 11 to 15 volts. The measured voltage should change by less than 0.2 V DC. Reset the power supply voltage to 14 V DC.

BAND VCO ADJUSTMENT

Measure the voltage at the point labeled 160-15 M EL (near C138) on the L.O. board. Adjust the coils according to the following chart for 3.75 ±0.25 V DC.

BAND	COIL
160M	T101
80M	T102
40M	T103
20M	T104
15M	T105

Measure the voltage at the point labeled 10 M EL on the LO board. Set the BAND switch to 10A and adjust T106 for 3.20 ± 0.1 V DC. Switch to 10B, 10C, and 10D and note that voltage reading increases approximately 0.4 V DC for each band.

BAND REFERENCE OSCILLATOR ADJUSTMENT

Connect a frequency counter to collector of Q111 on the LO board. Set bandswitch to 10D. Adjust the 2 MHz oscillator trimmer capacitor (C246) for a frequency of 44.000 MHz ±20 Hz. Disconnect the frequency counter.

CARRIER OSCILLATOR ADJUSTMENT

Connect the microphone to the MIC/KEY jack on the front panel. Connect the oscilloscope probe to the emitter of Q117 on the exciter board. Key the unit with the microphone button and adjust transformer T105 on the exciter board for a maximum indication on the oscilloscope of approximately 1 V p-p. Unkey the unit. Disconnect the oscilliscope probe and reconnect the frequency counter to the same point. Key the unit and make adjustments according to the following table, in the order listed.

MODE	CAPACITOR	FREQUENCY
LSB	C194	9.00000 MHz ±10 Hz
USB	C197	9.00330 MHz ±10 Hz
CW	C199	9.00250 MHz ±10 Hz

Unkey the unit and disconnect the frequency counter and microphone.

PASSBAND TUNING OSCILLATOR ADJUSTMENT

Measure the voltage at the center leg of the PASSBAND tuning control (R103B, rear section) and adjust the PASSBAND control for 5.25 ± 0.05 V DC.

Connect the frequency counter to G2 of Q113 on the exciter board and adjust the PBT trimmer capacitor (C206) for a frequency of 22.80165 MHz ± 20 Hz. Vary the PASSBAND control from fully counterclockwise to fully clockwise and note that the PBT oscillator has a frequency range greater than 3.6 kHz.

Disconnect the frequency counter and reconnect the oscilloscope probe to the same point. Adjust coil T106 for maximum indication on the oscilloscope screen, approximately 1.5 V p-p. Disconnect the oscilloscope probe.

COUNTER REFERENCE OSCILLATOR ADJUSTMENT

Remove the cover from the units frequency counter and connect the shop frequency counter to pin 9 of U114 on the counter board. Adjust the reference oscillator trimmer (C110) for a frequency of 1.31072 MHz ± 2 Hz. Disconnect the frequency counter and reinstall the counter cover.

PASSBAND TUNING INDICATOR ADJUSTMENT

Set the PASSBAND control to its 12 o'clock position and adjust the A-D RANGE trimpot (R166) on the audio board so that all eight PASSBAND KHZ lights are on.

PTO AND RIT ADJUSTMENT

Set the MODE switch to LSB, and the BAND switch to 20, turn the RIT to its off position (RIT light extinguished) and adjust PTO A for a display frequency of 14270.0. Switch the MODE switch to USB, and if the display frequency changes, adjust the USB SHIFT trimmer (C608) on PTO A so that there is no change in the display frequency when shifting between LSB and USB modes.

Set the RIT control to 0. Push the RIT control to turn it on (RIT light lighted). If the display frequency changes, adjust the RIT CENTER trimpot (R536) on the PTO control board so that there is no change in the display frequency between RIT on and RIT off. Verify that the RIT control varies the frequency of the display ± 800 to 1500 Hz.

Turn the PTO A control fully counterclockwise and note how much the display frequency goes below 14000.0 kHz. Now turn the PTO A control to its clockwise limit and note how much the display frequency goes above 14500.0 kHz. The amount that the frequency extends beyond the lower and upper band limits should be greater than 50 kHz and within 10 kHz of each other. If this is not the case, adjust trimmer C606 on PTO A for the proper frequency range.

Set the PTO MODE switch to PTO B and repeat the above procedures for B.

EXCITER I.F. ADJUSTMENTS

Set the MODE switch to CW, the PTO MODE switch to PTO A, the PASSBAND control to its 12 o'clock position, and adjust PTO A for a display frequency of 14300.0 kHz. Connect the output of the RF generator to the ANTENNA connector on the rear apron of the unit. Set the RF generator output for 50 uV and adjust its frequency for maximum S meter reading on the unit. Adjust transformers T101, T102, T103 and T104 on the exciter board for a maximum S meter reading.

AGC AND S METER CALIBRATION

Use the same test setup as previous section. Connect the oscilloscope probe to the cathode of diode CR123 (AF in line) on the audio board. Adjust the AGC THRESH trimpot (R101) on the audio board for 600 mV p-p indication on the oscilliscope screen. Disconnect the oscilloscope probe.

Adjust the S METER CAL trimpot (R110) on the audio board for an S 9 indication on the unit's meter.

CWN FILTER ADJUSTMENT

Set the MODE switch to CWN and connect the frequency counter across the units speaker terminals, keeping the rest of the test setup the same as for the previous section. Adjust PTO A for a 1650 Hz ± 20 Hz speaker audio tone and adjust CWN trimmer capacitors C130 and C134 on the exciter board for a maximum S meter reading. Switch the MODE to CW and verify that the S meter reading changes by less than one S unit.

NOTCH FILTER ADJUSTMENT

Use the same test setup as before. Connect the AC voltmeter or oscilloscope across the speaker terminals. Adjust PTO A for a 1650 ± 20 Hz tone in the speaker. Adjust the NOTCH FILT control for minimum speaker audio output, then adjust the NOTCH FILTER ADJUST trimpot (R178) on the audio board for minimum speaker audio output. Repeat the adjustments of NOTCH FILT and N F ADJ controls until there is no further decrease in the speaker audio output. The notch filter should attenuate the speaker audio by more than 25 dB. Verify that speaker audio output can also be attenuated by the notch filter with speaker audio tones of 400 Hz and 2500 Hz. Disconnect the frequency counter, AC voltmeter or oscilloscope and RF generator from the unit.

ALC AND METER ADJUSTMENTS

Connect the microphone to the MIC/KEY jack on the front panel of the unit. Connect the RF wattmeter and 50 ohm dummy load to the ANTENNA connection on the rear apron. Set the MODE switch to CW and adjust PTO A for a display frequency of 14500.0 kHz.

Connect the DC oscilloscope probe to the "high" potential leg of the REFL trimpot (R107) on the preselector board. Key the unit with the microphone button and adjust the MIC GAIN control for 100 watts of output. Quickly adjust the SWR NULL trimmer capacitor (C103) on the SWR/wattmeter board for minimum DC voltage on the oscilloscope (< 0.5 V DC).

NOTE: During this and subsequent CW tests, limit the "key down" time to 30 seconds or less to prevent overheating and resultant damage to the power amplifier devices. Allow time between transmissions to permit them to cool.

Disconnect the oscilloscope probe. Set the METER switch to the FWD position. Key the unit, and adjust the MIC GAIN for 100 watts of output and quickly adjust the FWD trimpot (R147) on the distribution board for a reading of 100 on the unit's meter.

Set the MIC GAIN control fully clockwise, key the unit, and quickly adjust the FWD ALC trimpot (R104) on the preselector board for 110 watts of output.

Disconnect the 50 ohm dummy load and reconnect the 100 watt (10 ohm cold) light bulb in its place. With the MIC GAIN control fully clockwise, key the unit and quickly adjust the REFL ALC trimpot (R107) on the preselector board for 50-60 watts of output. Set the METER switch to REFL, key the unit, and quickly adjust the REFL trimpot (R149) on the distribution board for a reading of 100 on the unit's meter. (10 watts reflected power)

Check the output on all bands for 40 to 70 watts of output using the light bulb as a load.

Disconnect the light bulb load and reconnect the 50 ohm dummy load. Check the power output for 100-120 watts on all bands, with the MIC GAIN control fully clockwise.

Disconnect the coax lead at the ANTENNA connector of the unit. Key the unit with the MIC GAIN control fully clockwise and note that the total current drawn from the power supply is less than 12 amperes on all bands. Reconnect the coax lead to the ANTENNA connector.

NOTE: It may be necessary to slightly readjust the FWD and REFL ALC trimpots and SWR NULL trimmer capacitor to obtain the proper indications on all bands.

Set the BAND switch to 20 meters, the METER switch to ALC and the MIC GAIN control fully counterclockwise. Key the unit and adjust the ALC ZERO trimpot (R110) on the preselector board for a reading of S 2 on the units meter. Adjust the MIC GAIN control to its fully clockwise position. The meter reading should increase to +20 or greater. Unkey the unit.

PRESELECTOR AND L.O. BANDPASS FILTER ADJUSTMENT

The alignment of the bandpass filters should be performed only if absolutely necessary, and then only for those bands that require alignment. Indications of misalignment are significant reduction in power output and receiver sensitivity that is not the result of a component failure or misalignment of transformers, coils and/or oscillators on the exciter board.

If alignment is to be performed using a sweep generator, follow the procedure supplied with the test equipment and test the boards as follows:

Preselector Bandpass Filters: (with sweeper)

Connect the sweep generator output to the RX IN jack (J132) and the spectrum analyzer's input to the RX OUT jack (J136). Sweep each band according to the accompanying chart.

L.O. Bandpass Filters: (with sweeper)

Disconnect the leg of C177 from pin 7 of Z101 on the LO board and connect the sweep generator output to the leg of C177. Connect the analyzer input to the LO OUT jack (J103).

Sweep each band according to the accompanying chart.

Preselector Bandpass Filters: (without sweeper)

Disconnect the TX OUT coax at J134 on the preselector board and connect the oscilliscope probe to J134. Set the MODE switch to CW, and MIC GAIN control to fully clockwise, and key the unit. Adjust the oscilloscope for a reasonable indication.

Adjust the bandpass coils for maximum indication on the oscilloscope at the center frequency of each band.

Tune the PTO across each band and readjust the bandpass coils as necessary to obtain a flat response (± 2 dB) across the band without significantly reducing the maximum output.

L.O. Bandpass Filters: (without sweeper)

Disconnect the LO OUT coax at J103 on the LO board and connect the oscilliscope probe to J103. Use the same procedure for adjusting these bandpass coils as in the preceeding section.

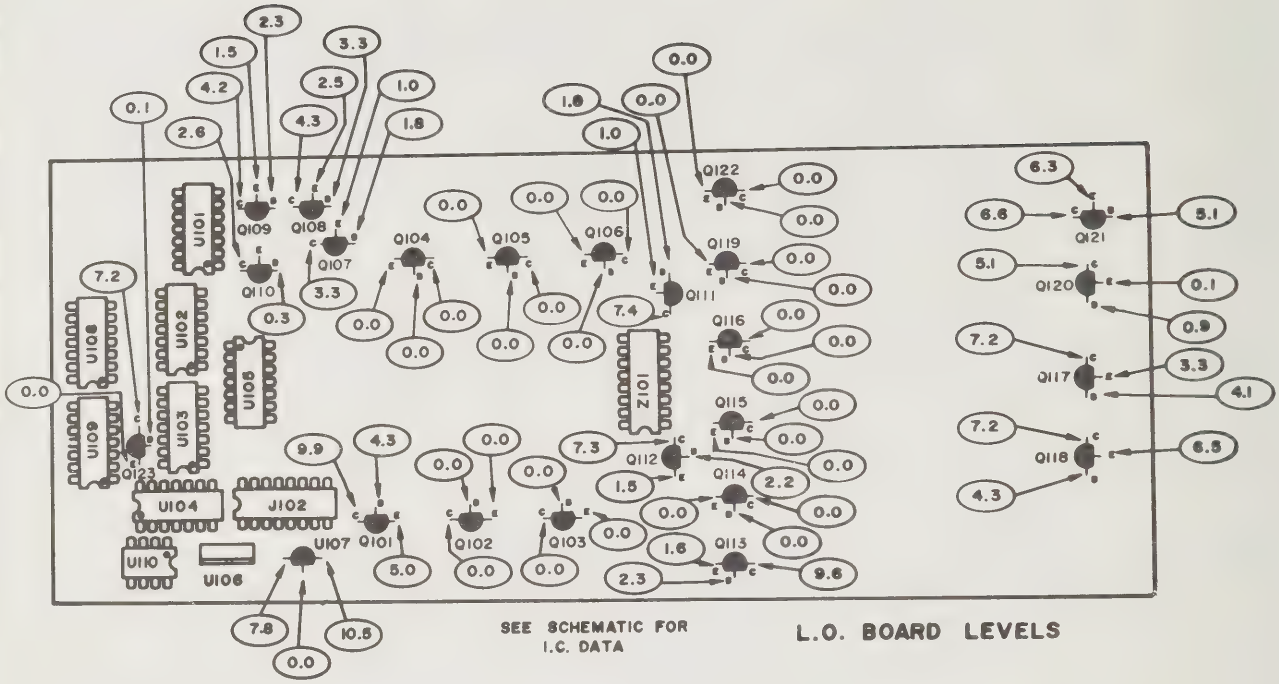
NOTE: It is not necessary to key the unit for LO bandpass filter adjustment.

PRESELECTOR BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHZ	CENTER FREQ.	COILS
160	1.8-2.0	1.9	T101, T102, T103
80	3.5-4.0	3.750	T104, T105, T106
40	7.0-7.5	7.250	T107, T108, T109
20	14.0-21.5	14.250	T110, T111, T112
15	21.0-11.5	21.250	T113, T114, T115
10A, B, C, D	28.0-30.0	29.0	T116, T117, T118

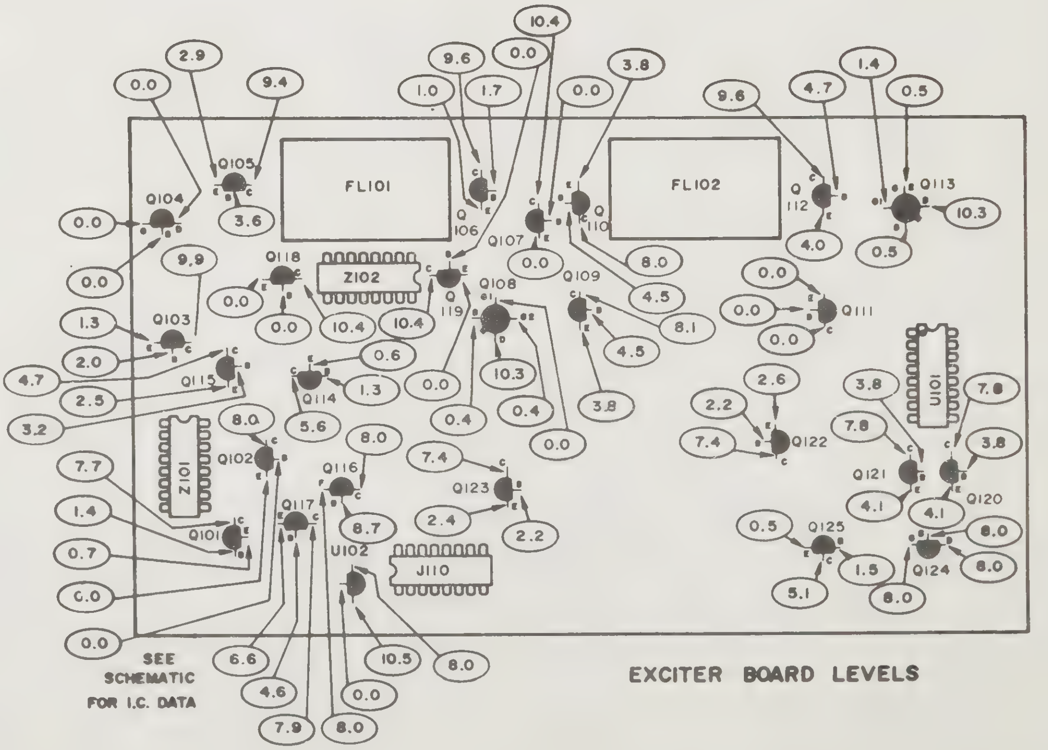
L.O. BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHZ	CENTER FREQ.	COILS
160	10.8-11.0	10.9	T107, T108, T109
80	12.5-13.0	12.750	T110, T111, T112
40	16.0-16.5	16.250	T113, T114, T115
20	23.0-23.5	23.250	T116, T117, T118
15	30.0-30.5	30.250	T119, T120, T121
10A, , C, D	37.0-39.0	38.0	T122, T123, T124



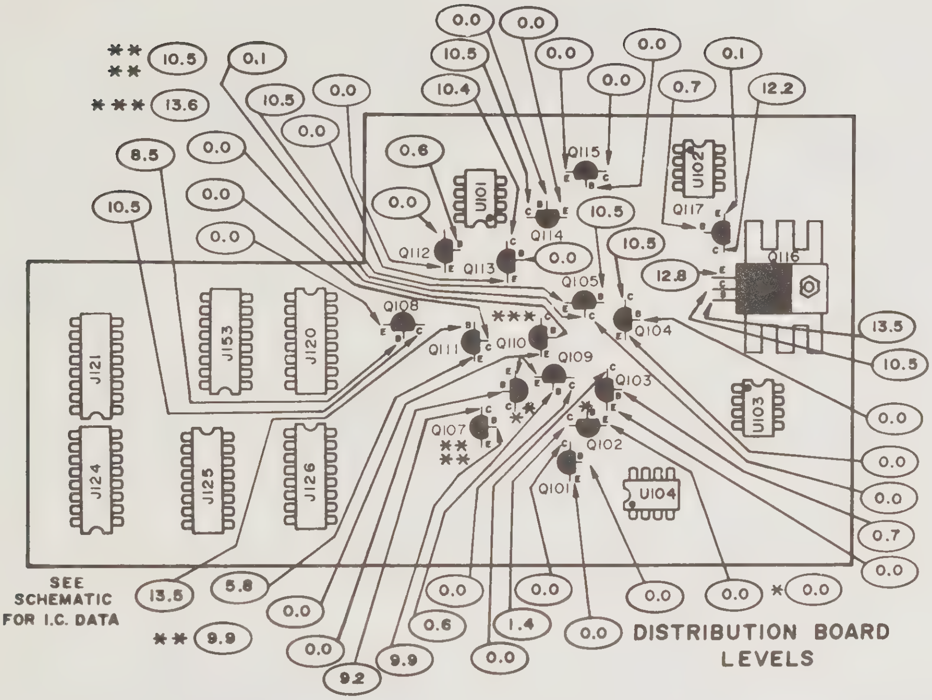
SEE SCHEMATIC FOR
I.C. DATA

L.O. BOARD LEVELS

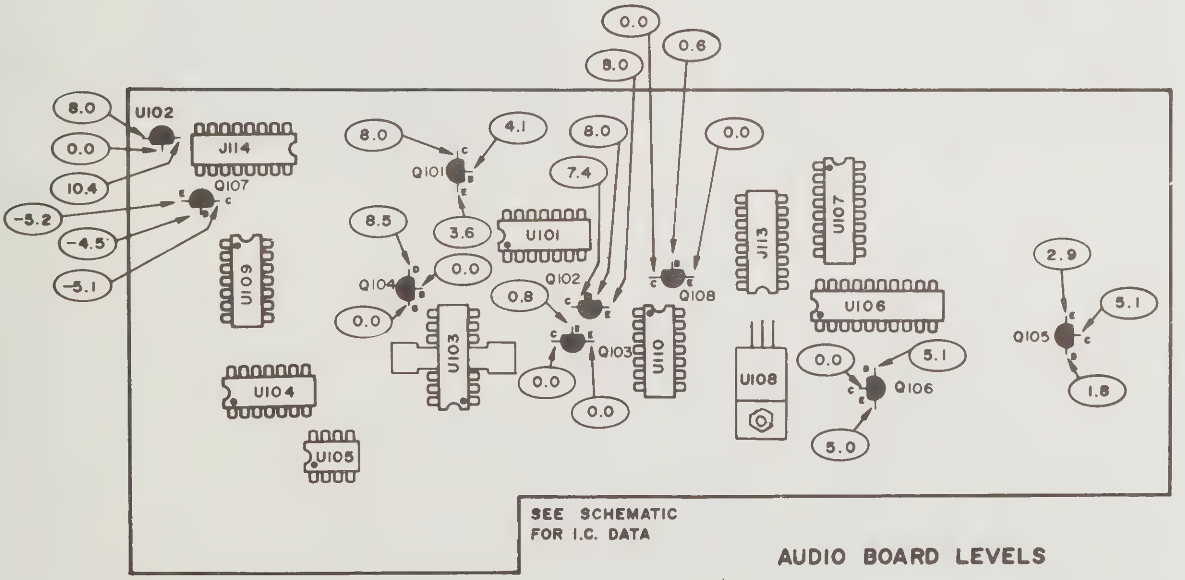


SEE
SCHEMATIC
FOR I.C. DATA

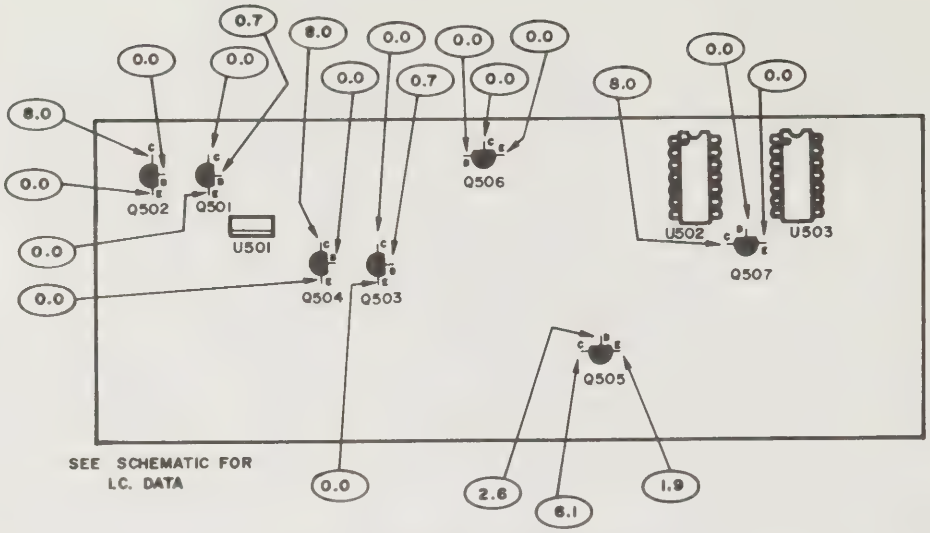
EXCITER BOARD LEVELS



DISTRIBUTION BOARD LEVELS

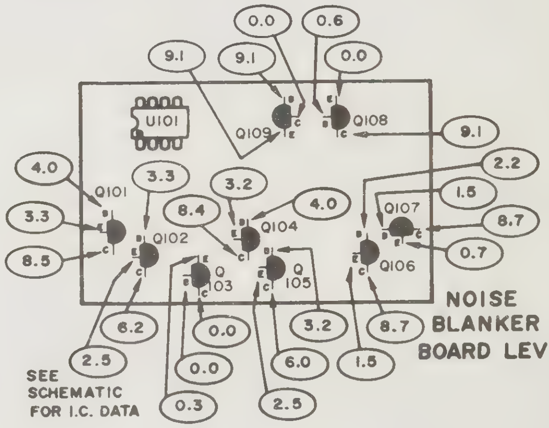


AUDIO BOARD LEVELS



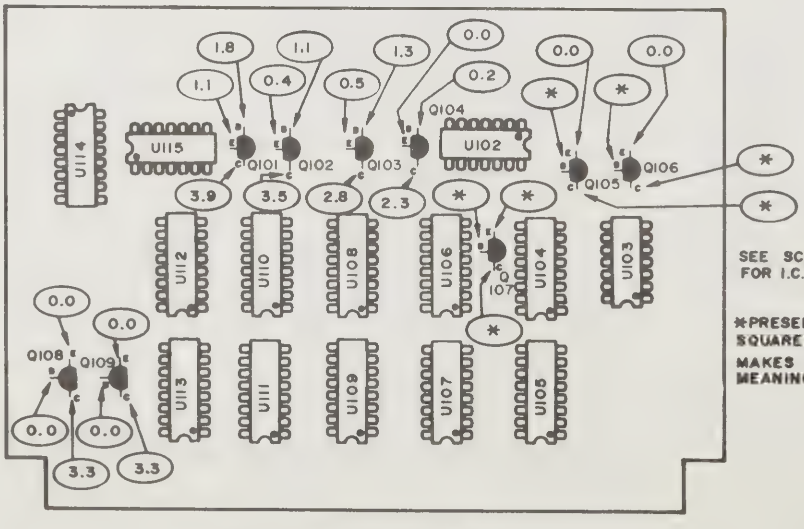
SEE SCHEMATIC FOR I.C. DATA

PTO CONTROL BOARD LEVELS



NOISE BLANKER BOARD LEVELS

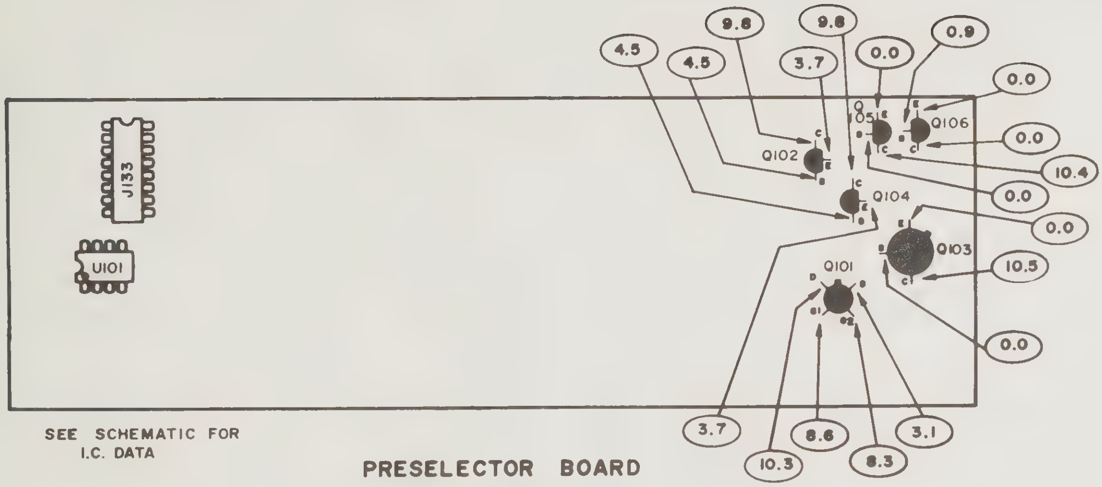
SEE SCHEMATIC FOR I.C. DATA



SEE SCHEMATIC FOR I.C. DATA

*PRESENCE OF SQUARE WAVE MAKES READING MEANINGLESS

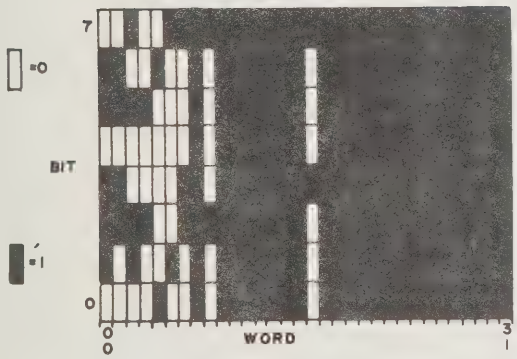
COUNTER BOARD LEVELS



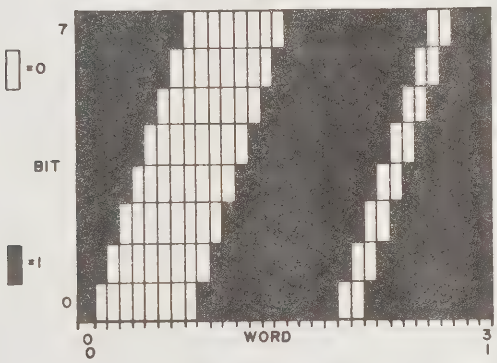
SEE SCHEMATIC FOR
I.C. DATA

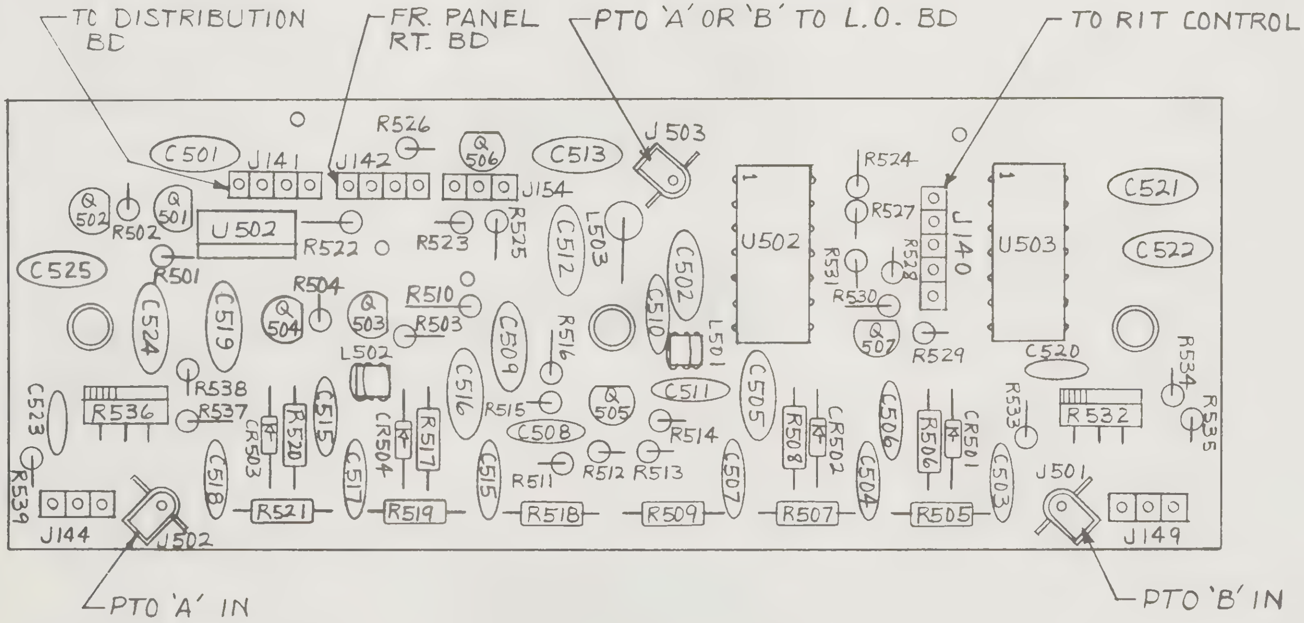
PRESELECTOR BOARD
LEVELS

U105 MEMORY MAP



U107 MEMORY MAP





SWAN ELECTRONICS

SCALE: ~

DATE: 9-18-79

APPROVED BY: JW

DRAWN BY AJG

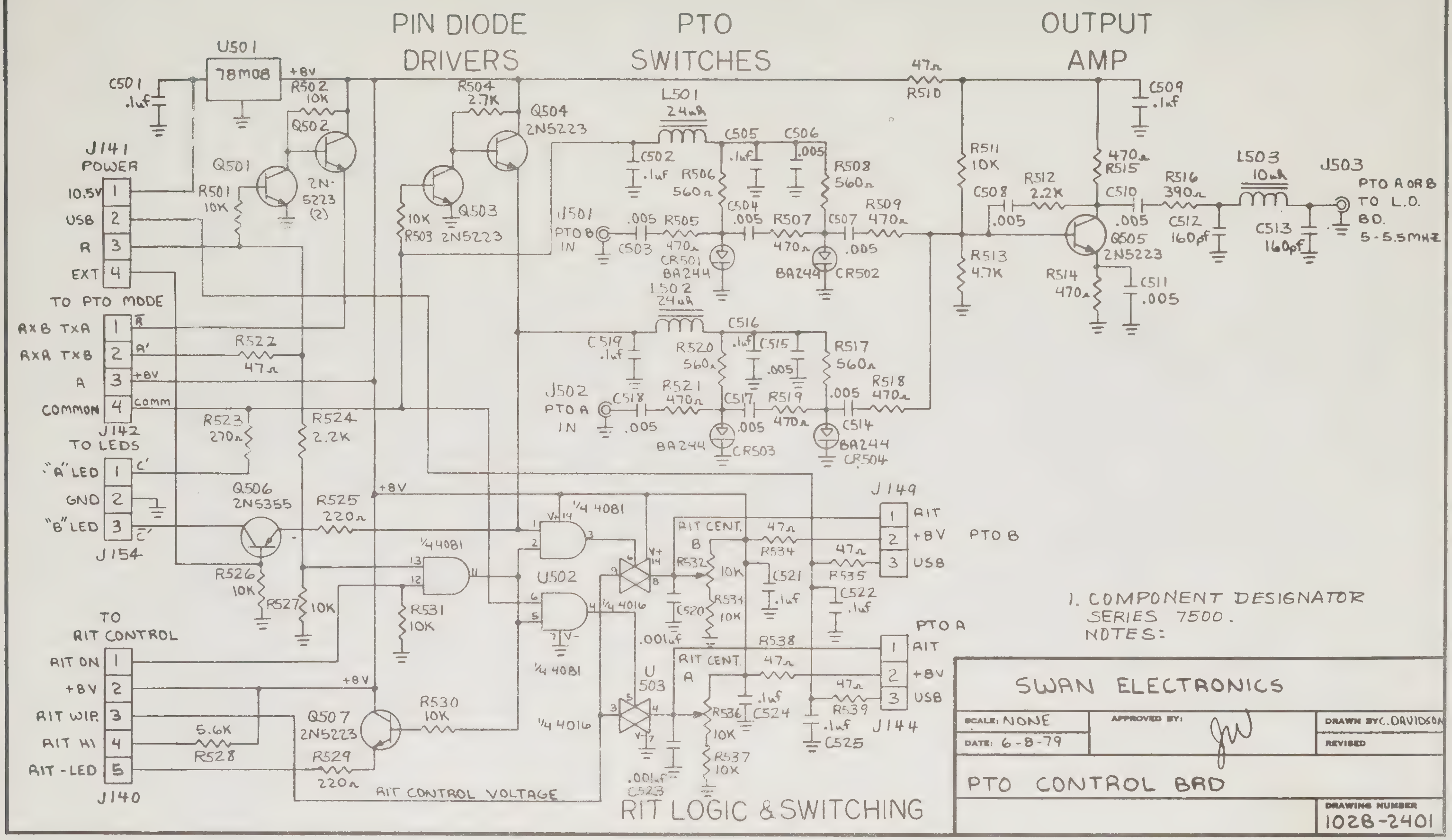
REVISED

PTO CONTROL BOARD LAYOUT

DRAWING NUMBER 102BX-2301

2. COMPONENT DESIGNATORS: 7500 SERIES
1. REF. ELEC. ASSY 102BX-2001
REF CIRCUIT BD 102BX-2101
REF DRILL DWG 102BX-2201
NOTES:

11X17 PRINTED ON NO. 1000H CLEARPRINT



1. COMPONENT DESIGNATOR SERIES 7500.
NOTES:

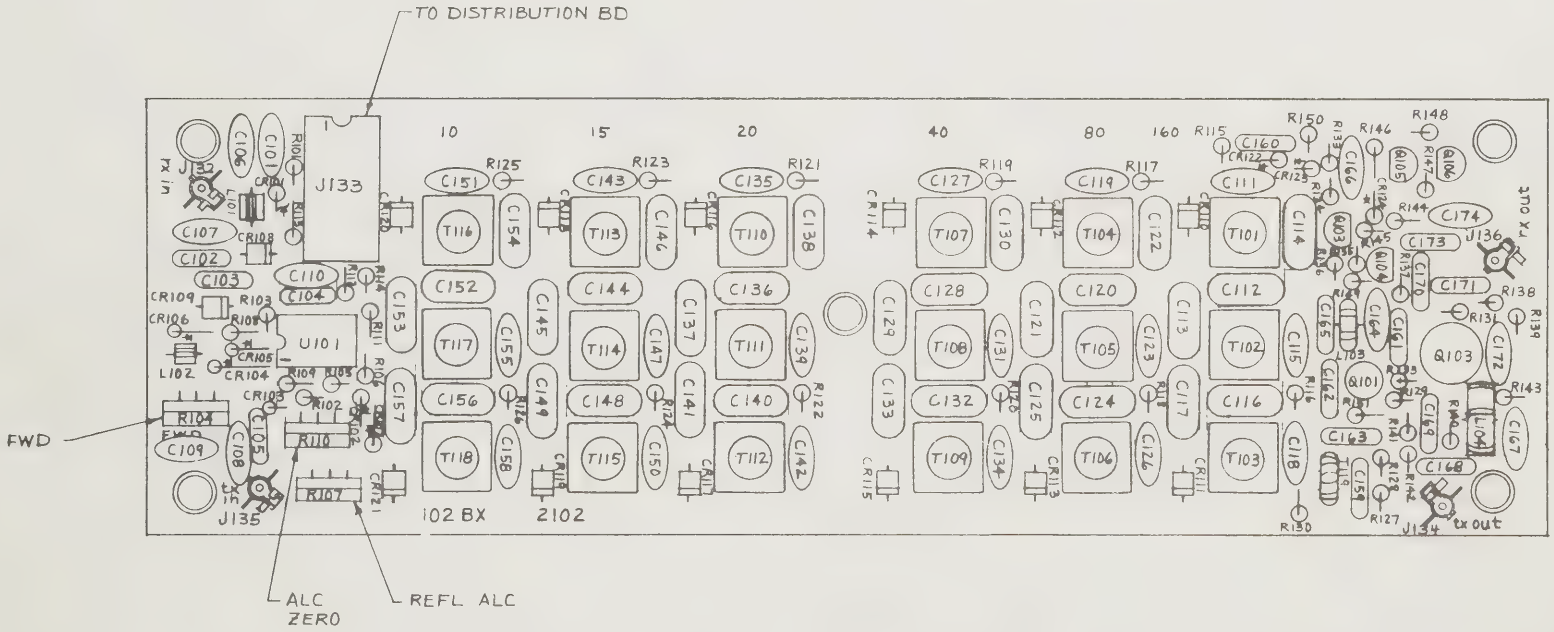
SWAN ELECTRONICS

SCALE: NONE
DATE: 6-8-79
APPROVED BY: JW
DRAWN BY: C. DAVIDSON
REVISED

PTO CONTROL BRD

DRAWING NUMBER 102B-2401

11X17 PRINTED ON NO. 1000H-10 CLEARPRINT FADE-OUT



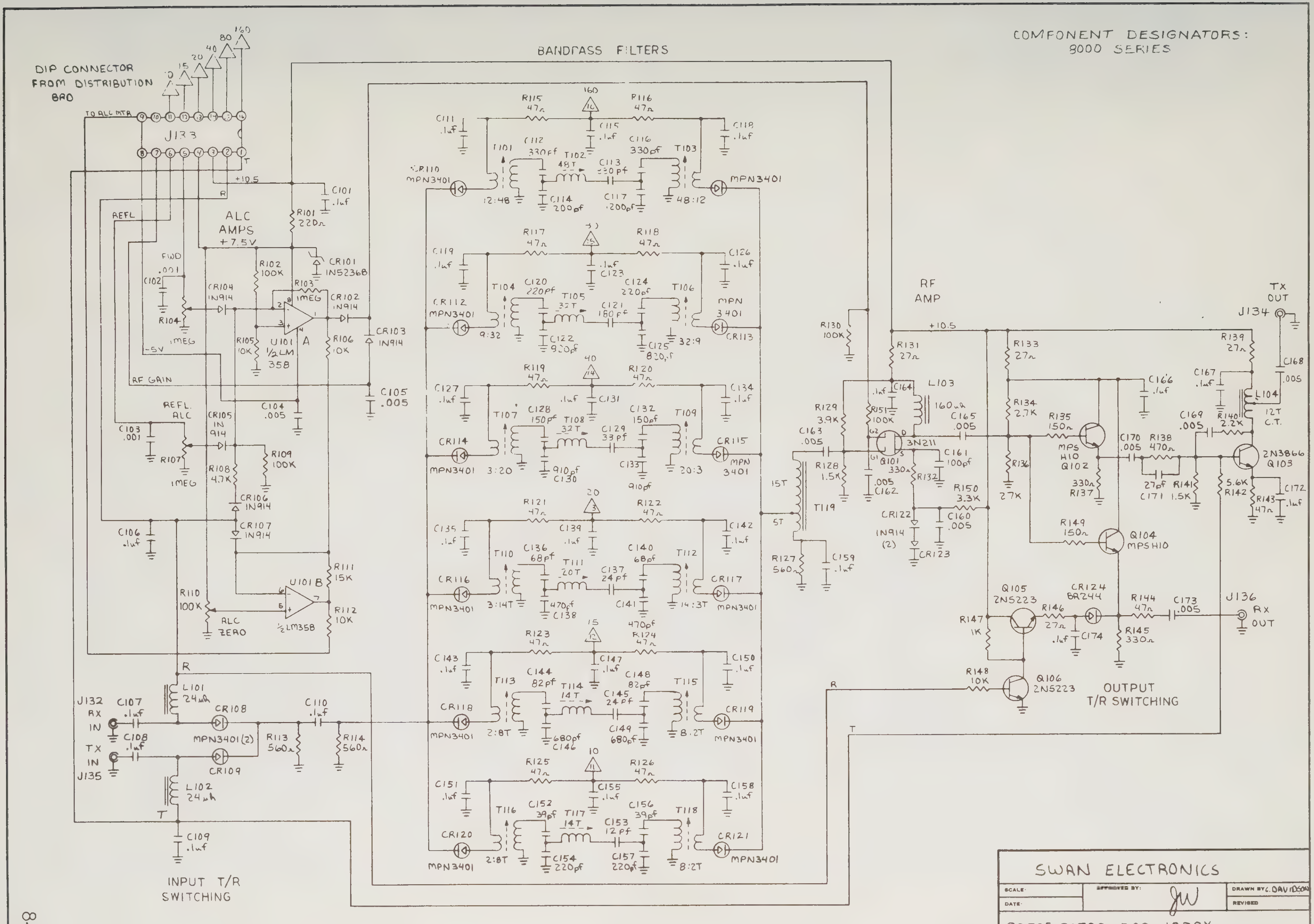
2. COMPONENT DESIGNATION SERIES 8000.

1. REF. ELECT. ASSY 102 BX- 2002
   REF. CIRCUIT BD 102 BX- 2102
   REF. DRILL DWG 102 BX- 2202

NOTES:

SWAN ELECTRONICS
SCALE: NONE
DATE: 7-24-79
APPROVED BY: JW
DRAWN BY: AJG
REVISED
PRESELECTOR BOARD
LAYOUT
REV
DRAWING NUMBER 102BX-2302

COMFONENT DESIGNATORS:
8000 SERIES



SWAN ELECTRONICS

SCALE:
DATE:
APPROVED BY: JW
DRAWN BY: C. DAVIDSON
REVISED

PRESELECTOR BRD 1028X

DRAWING NUMBER
1028-2402

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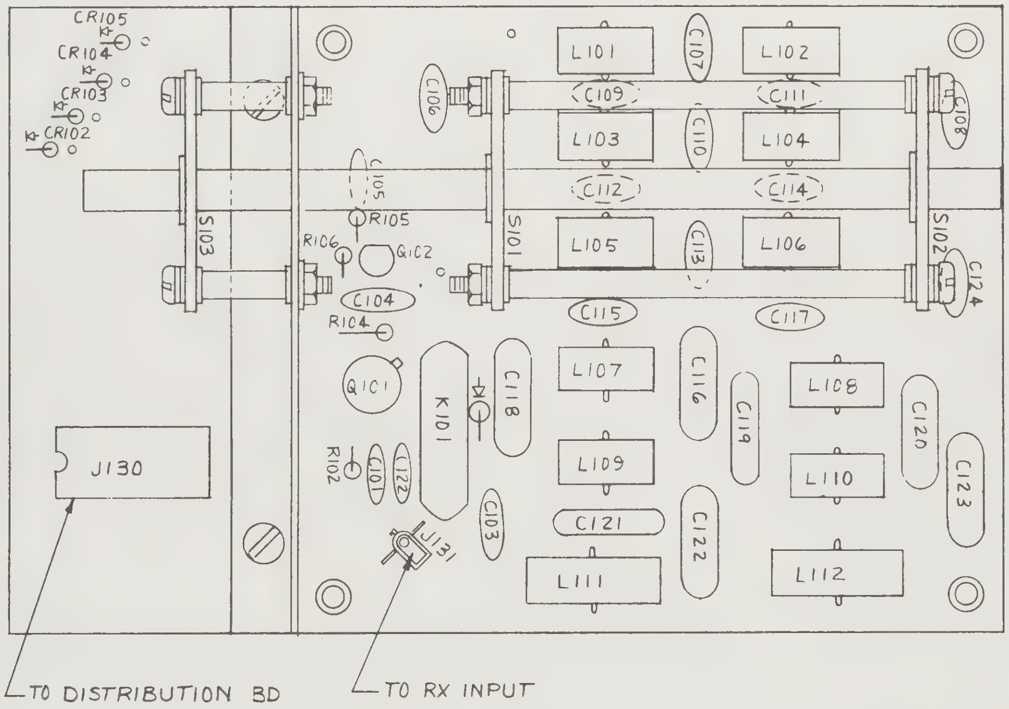
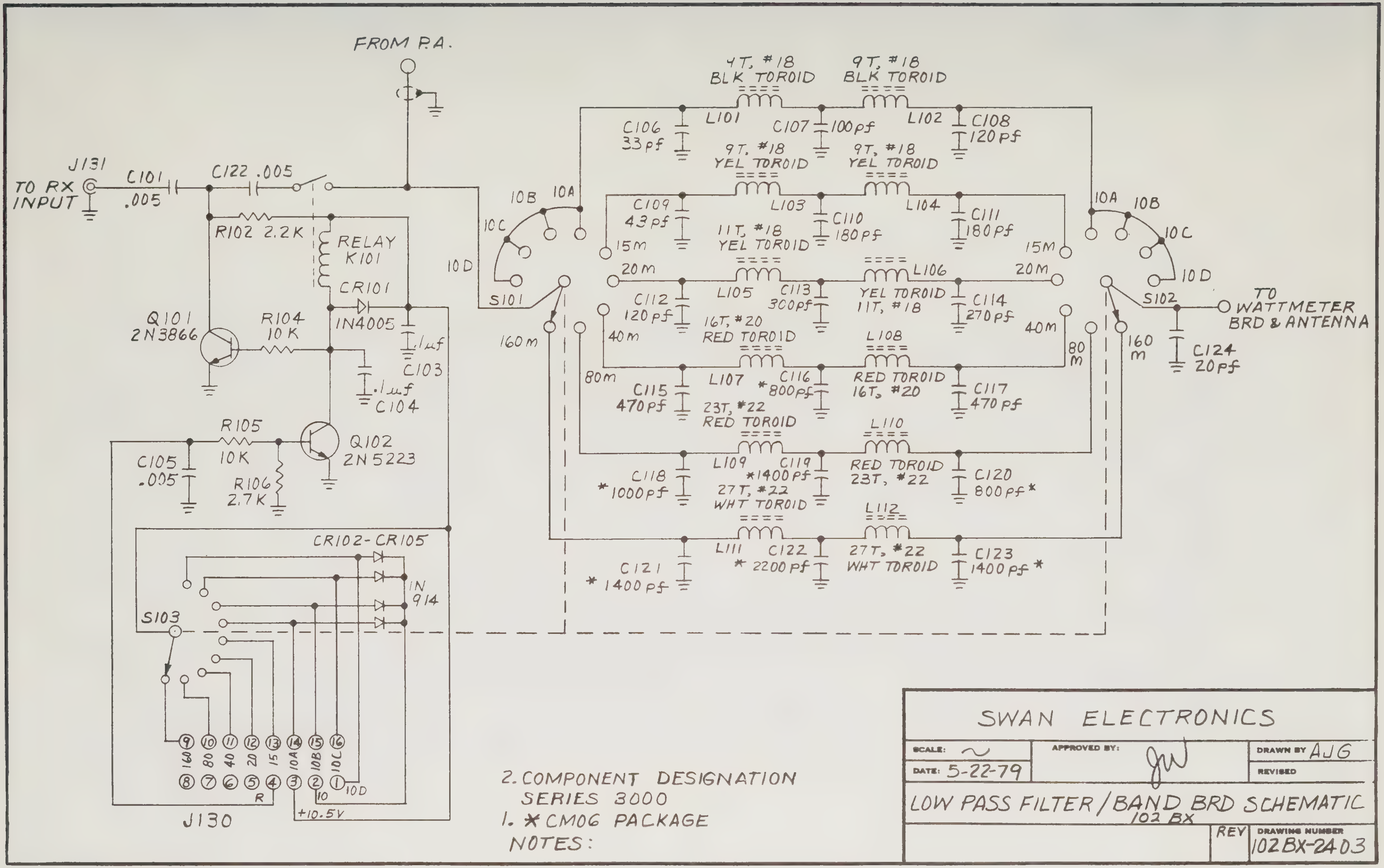


2. COMPONENT DESIGNATORS 3000 SERIES.
1. REF. ELEC. ASSY 102BX-2003
REF. CIRCUIT BD 102BX-2103
REF. DRILL DWG 102BX-2203
NOTES:

SWAN ELECTRONICS
SCALE: ~
APPROVED BY: JW
DRAWN BY AJG
DATE: 9-18-79
REVISED
B'SWITCH & L.P. FILT. BD LAYOUT
DRAWING NUMBER 102BX-2303



2. COMPONENT DESIGNATION SERIES 3000

1. * CM06 PACKAGE

NOTES:

SWAN ELECTRONICS

SCALE: ~

APPROVED BY: JW

DRAWN BY AJG

DATE: 5-22-79

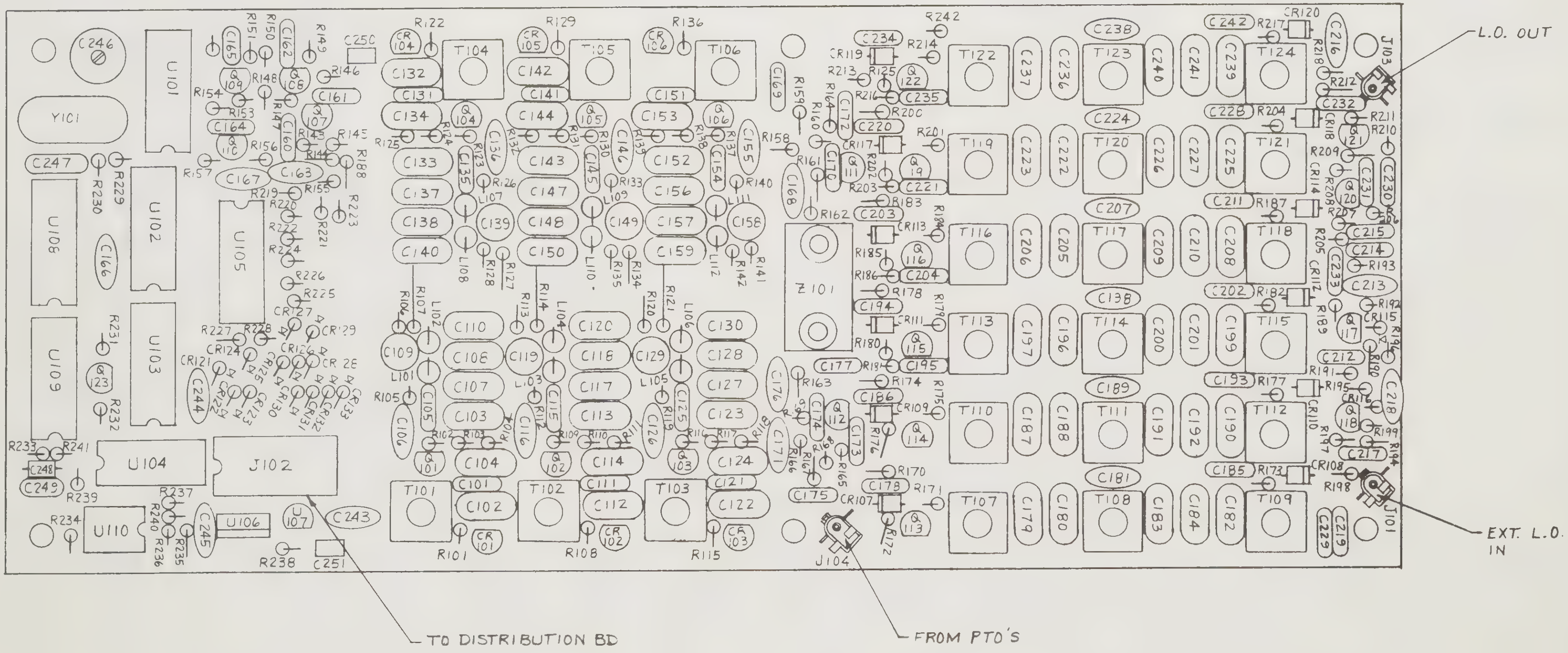
REVISED

LOW PASS FILTER / BAND BRD SCHEMATIC 102 BX

REV

DRAWING NUMBER 102BX-2403

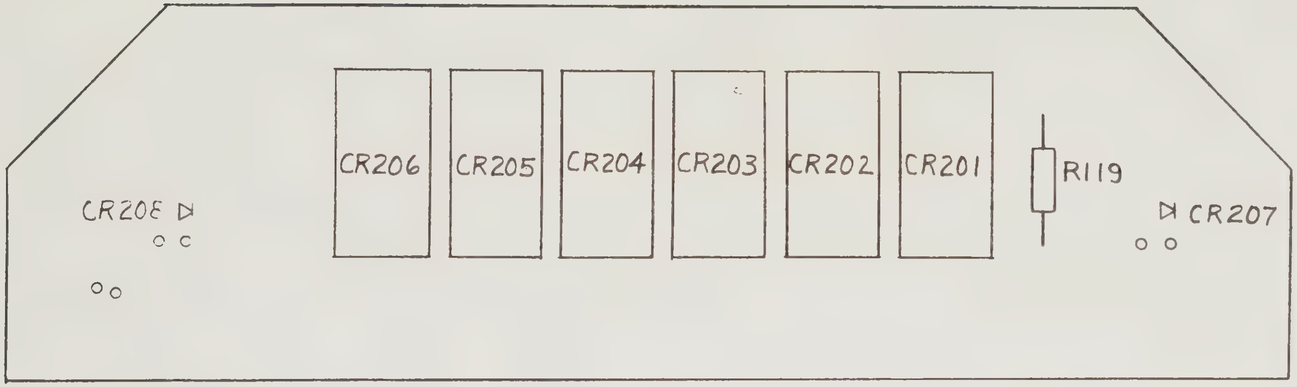
11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS: 5000 SERIES

1. REF. ELECT. ASSY 102BX-2004
REF. CIRCUIT BOARD 102BX-2104
REF. DRILL DWG 102BX-2204
NOTES.

SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 7-30-79		REVISED
L.O. BOARD COMPONENT LAYOUT		
	REV.	DRAWING NUMBER 102BX-2304



2. COMPONENT DESIGNATOR SERIES 7200
1. REF. ELECT. ASSY 102BX-2006
REF. CIRCUIT BD 102BX-2106
REF. DRILL DWG 102BX-2206
NOTES:

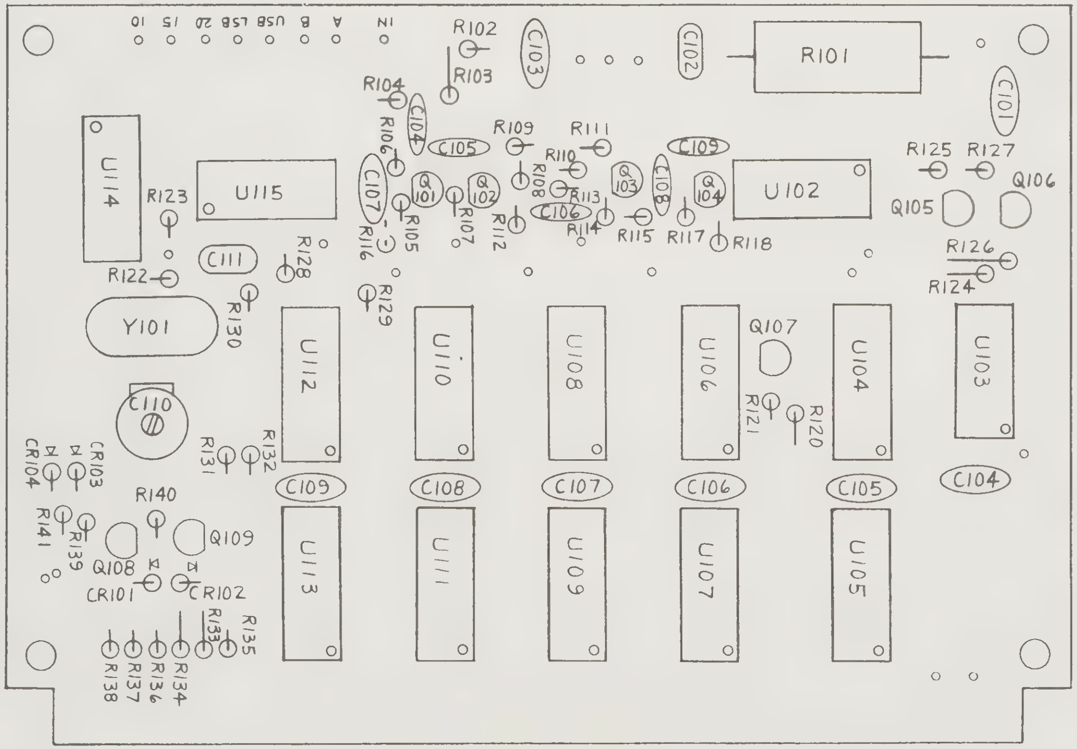
SWAN ELECTRONICS

SCALE: ~
DATE: 9-19-79
APPROVED BY: JW
DRAWN BY AJG
REVISED

DISPLAY BOARD LAYOUT

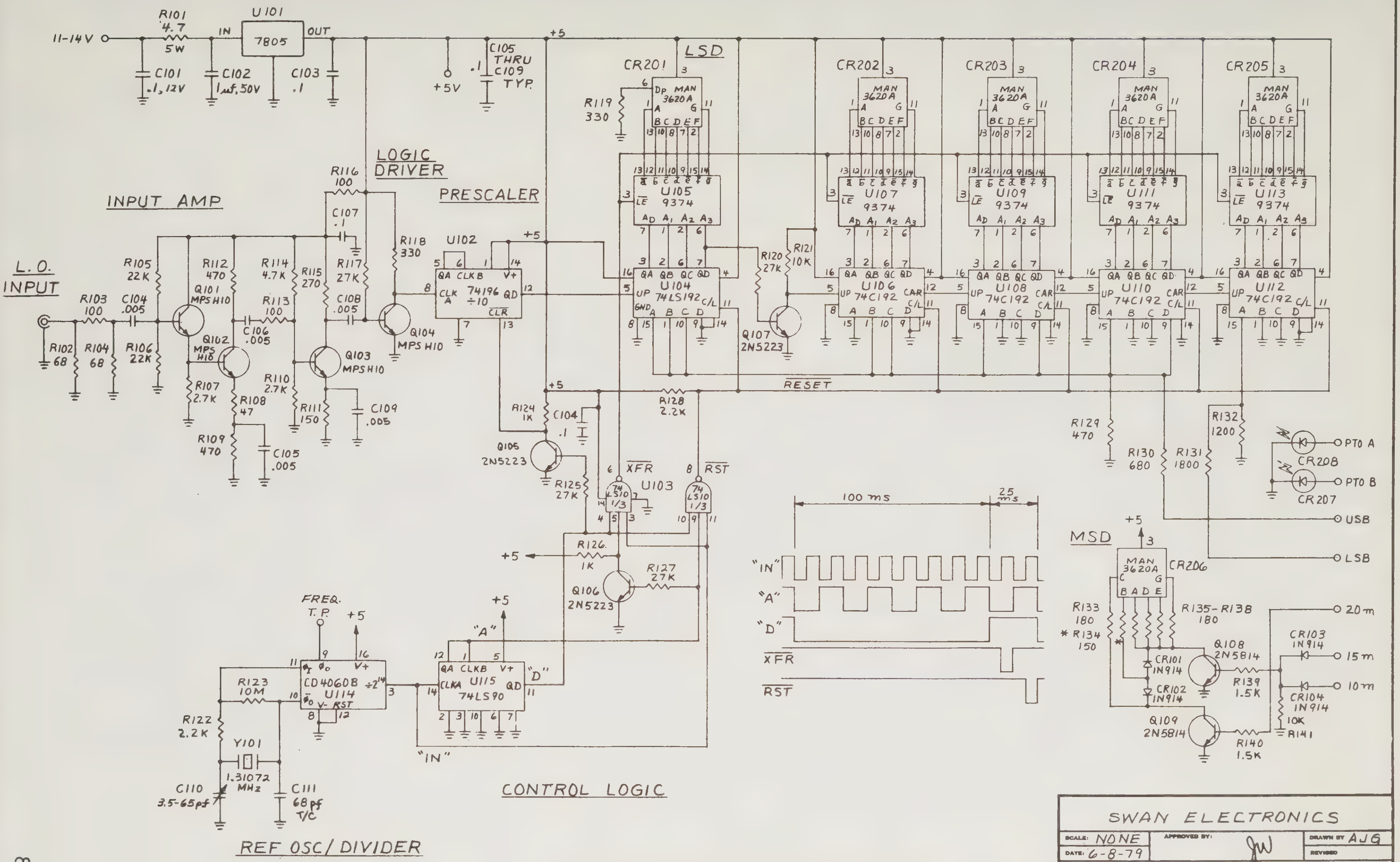
DRAWING NUMBER 102BX-2306

11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS 7100 SERIES
1. REF. ELEC. ASSY 102BX-2005
REF. CIRCUIT BD 102BX 2105
REF. DRILL DWG 102BX 2205
NOTES:

SWAN ELECTRONICS
SCALE: ~
APPROVED BY: JW
DRAWN BY AJG
DATE: 9-19-79
REVISED
COUNTER BOARD LAYOUT
DRAWING NUMBER 102BX-2305



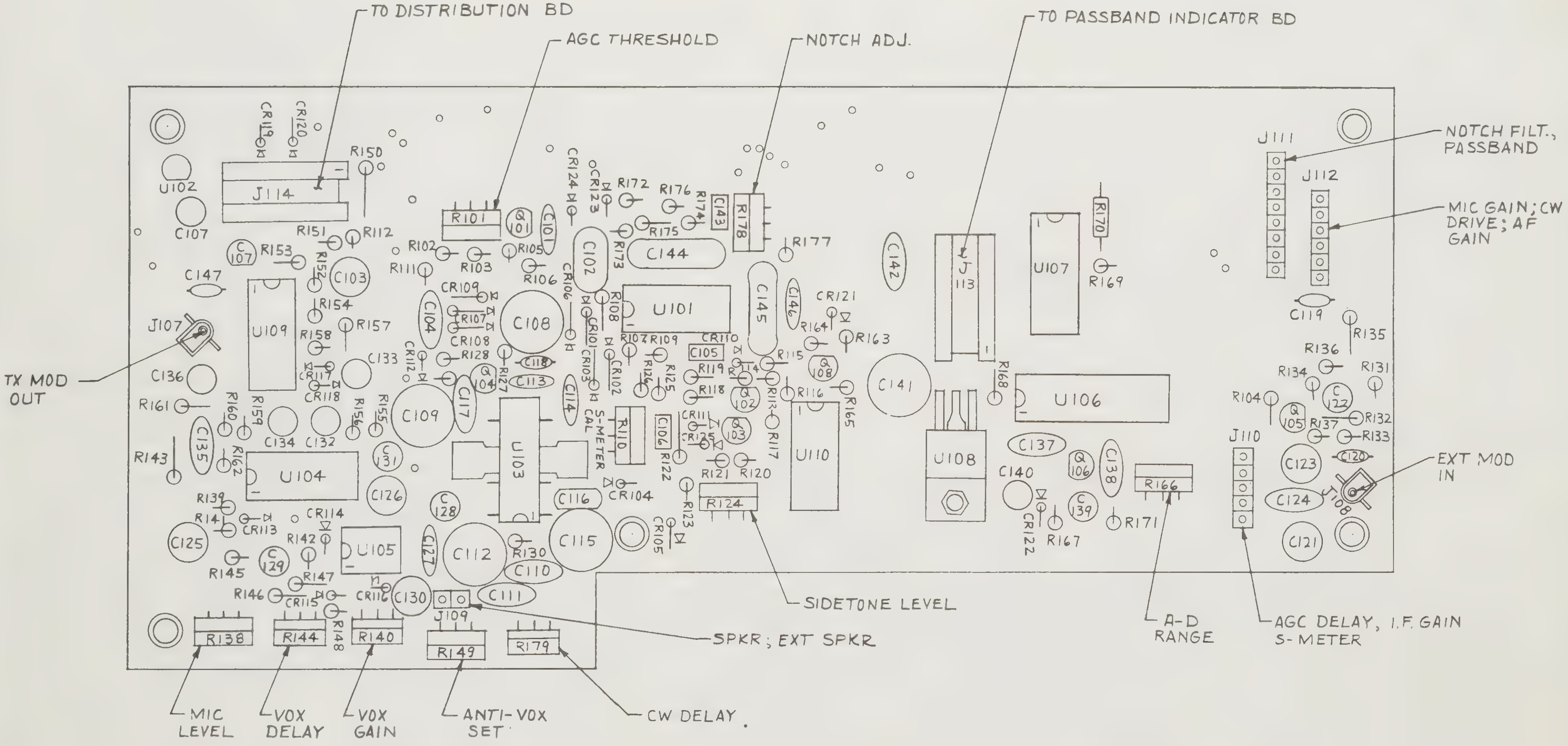
2. DISPLAY REF. DESIGNATOR SERIES 7200
1. COUNTER REF. DESIGNATOR SERIES 7100
NOTES:

SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 6-8-79		REVISED

COUNTER / DISPLAY BD SCHEMATIC
102 BX

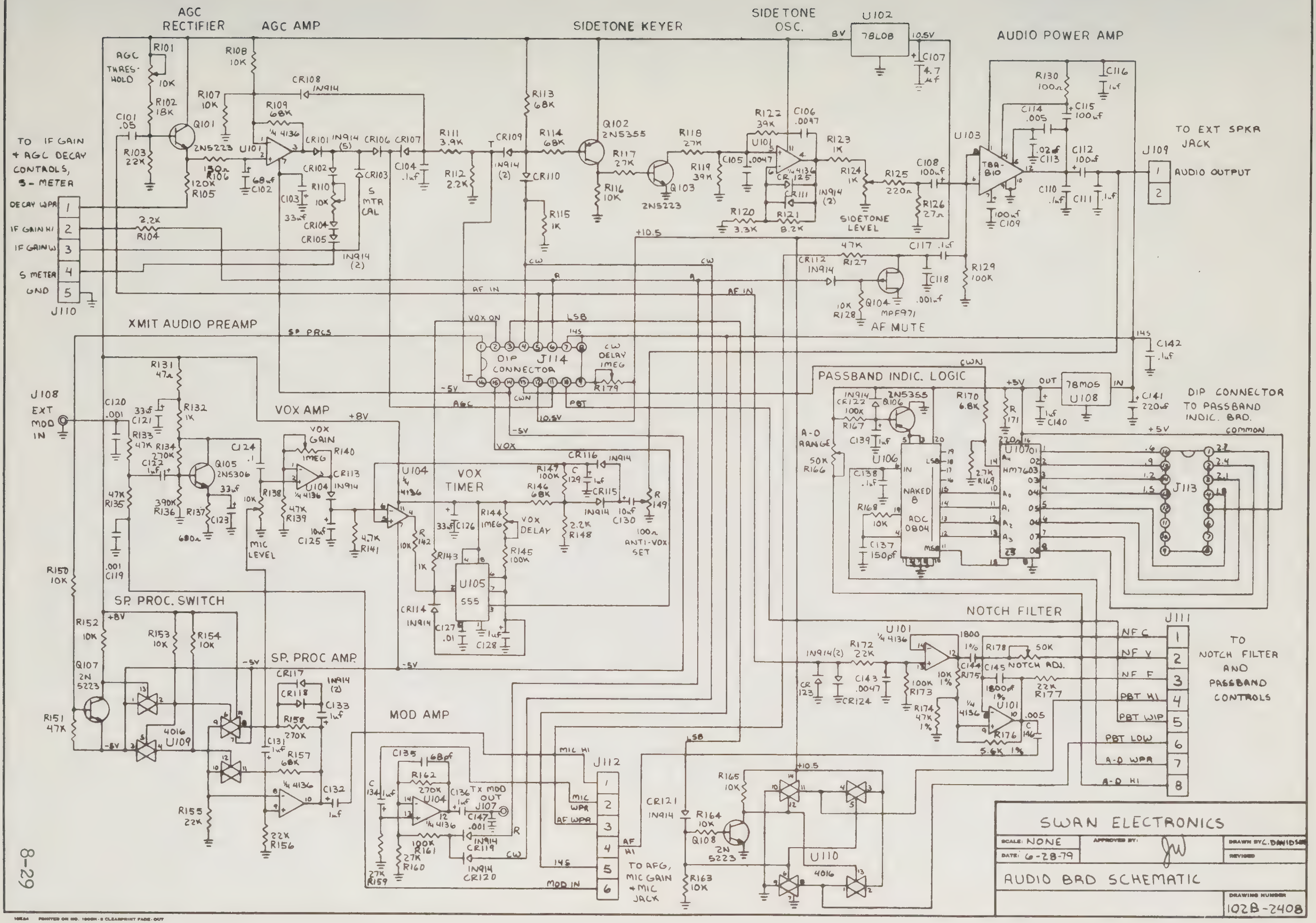
REV DRAWING NUMBER 102BX-2405



2. COMPONENT DESIGNATORS 4000 SERIES.
1. REF. ELECT. ASSY 102BX-2008
REF. CIRCUIT BOARD 102BX-2108
REF. DRILL DWG 102BX-2208
NOTES:

SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE:		REVISED
AUDIO BOARD LAYOUT		
		DRAWING NUMBER 102BX-2308

NOTES: REF. DESIGNATION SERIES 4000.



SWAN ELECTRONICS

SCALE: NONE

APPROVED BY: JW

DRAWN BY: C. DAVIDSON

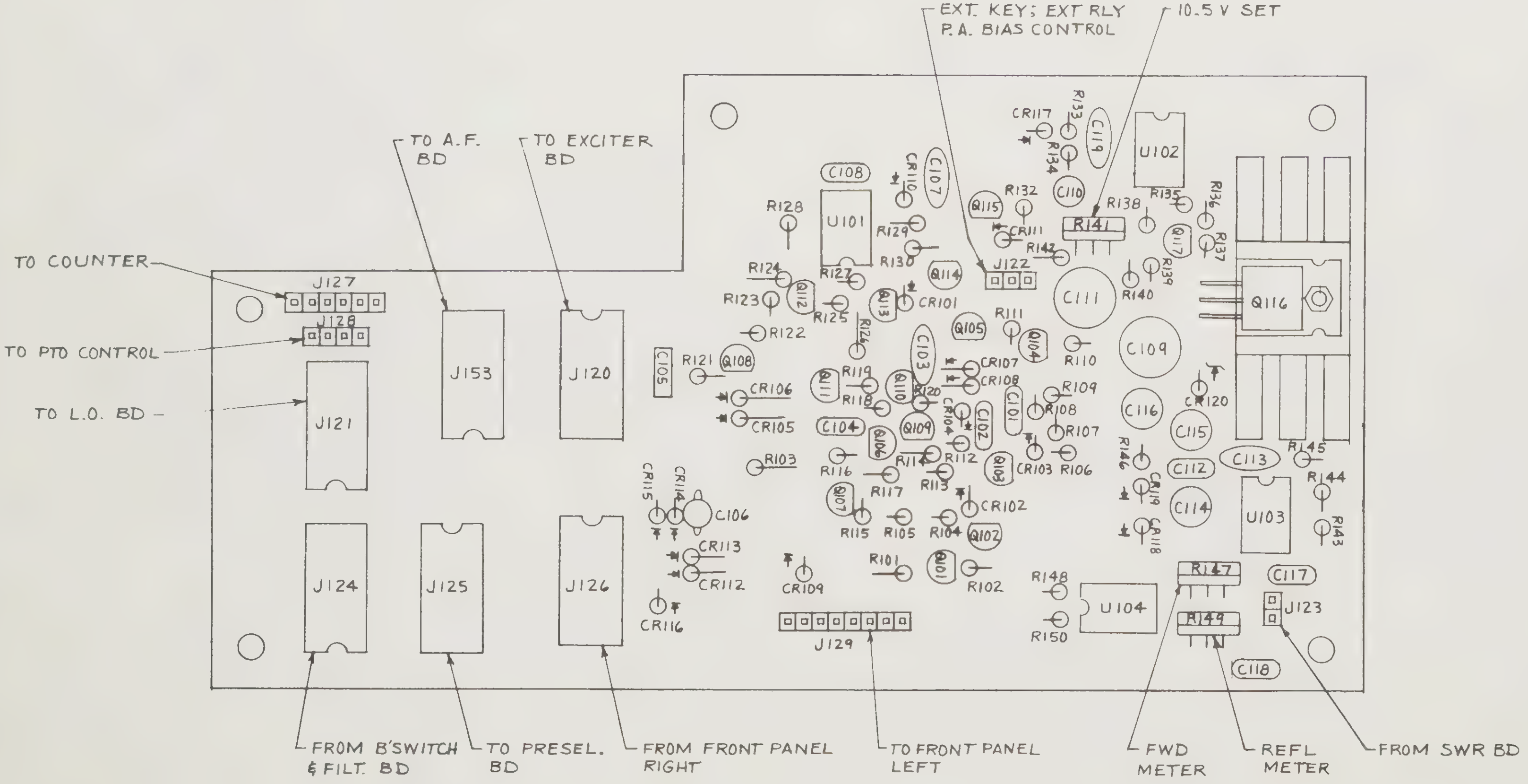
REVISED

DATE: 6-28-79

AUDIO BRD SCHEMATIC

DRAWING NUMBER 102B-2408

PRINTED ON NO. 1000H-5 CLEARPRINT FADE-OUT

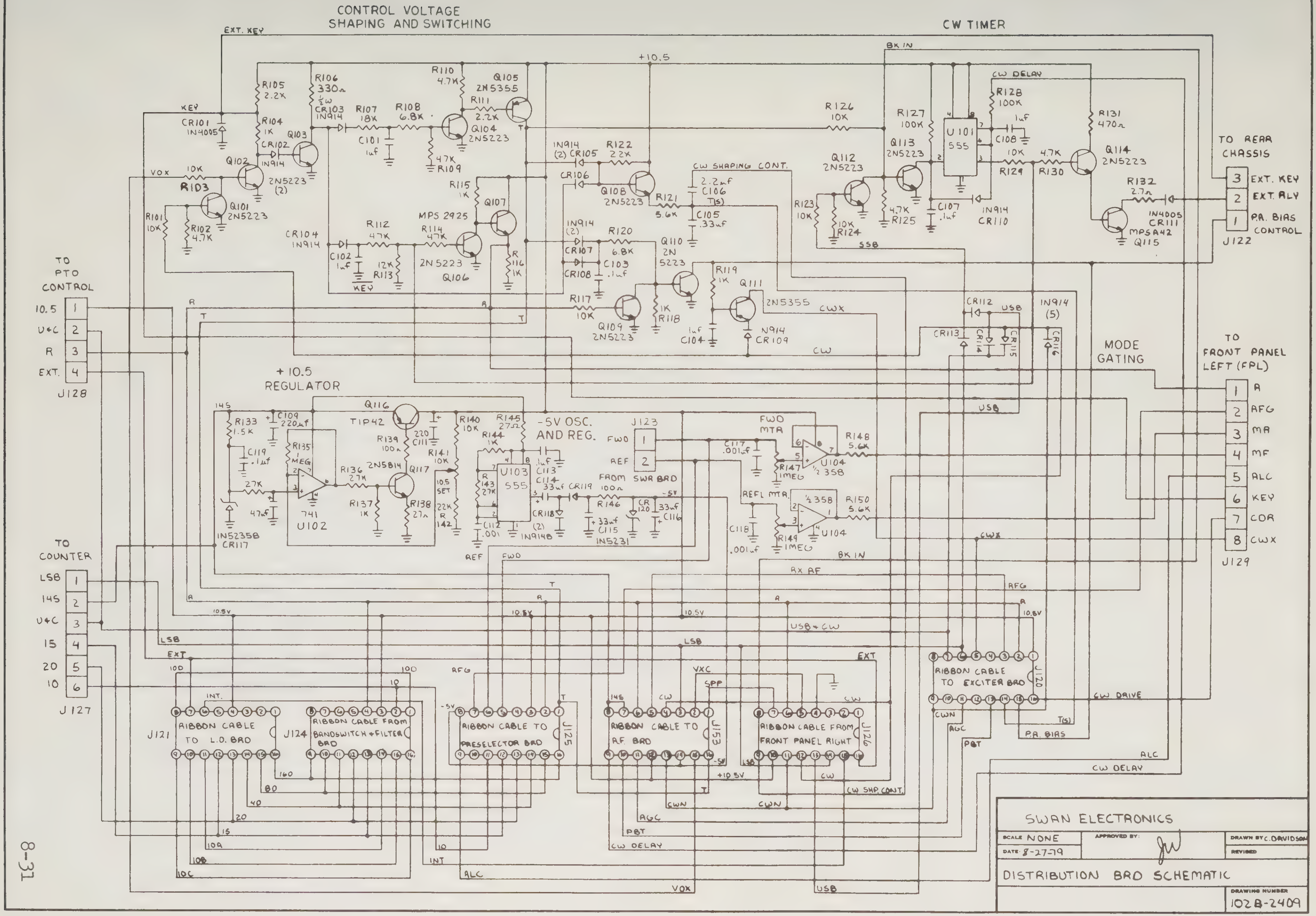


2. COMPONENT DESIGNATORS: 9000 SERIES

1. REF. ELECT. ASSY 102BX-2009
REF. CIRCUIT BD 102BX-2109
REF. DRILL DWG 102BX-2209
NOTES:

SWAN ELECTRONICS
SCALE: NONE
APPROVED BY: JW
DRAWN BY AJG
DATE 7-25-79
REVISED
DISTRIBUTION BOARD
COMPONENT LAYOUT
REV
DRAWING NUMBER 102BX-2309

COMPONENT DESIGNATORS:
9000 SERIES

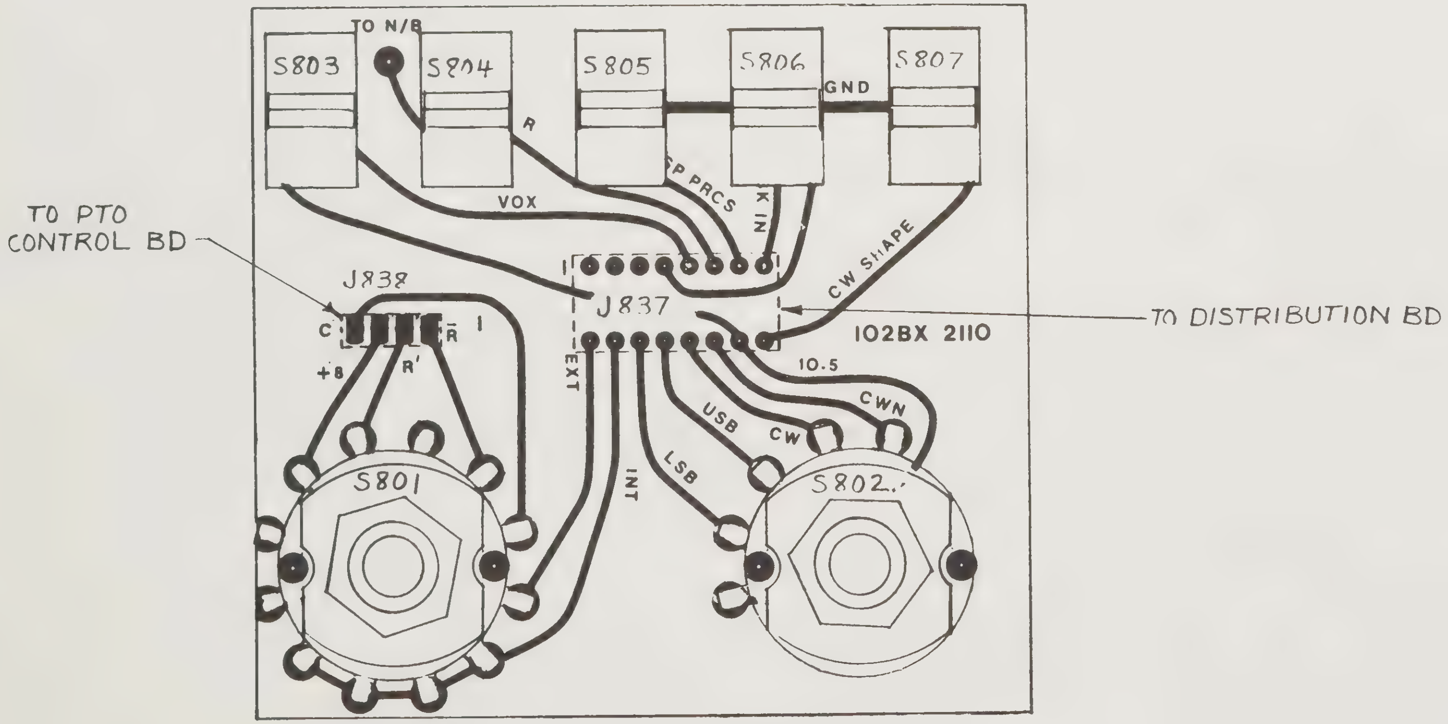


SWAN ELECTRONICS

SCALE NONE
DATE 8-27-79
APPROVED BY: JW
DRAWN BY C. DAVIDSON
REVISED

DISTRIBUTION BRD SCHEMATIC

DRAWING NUMBER 102B-2409



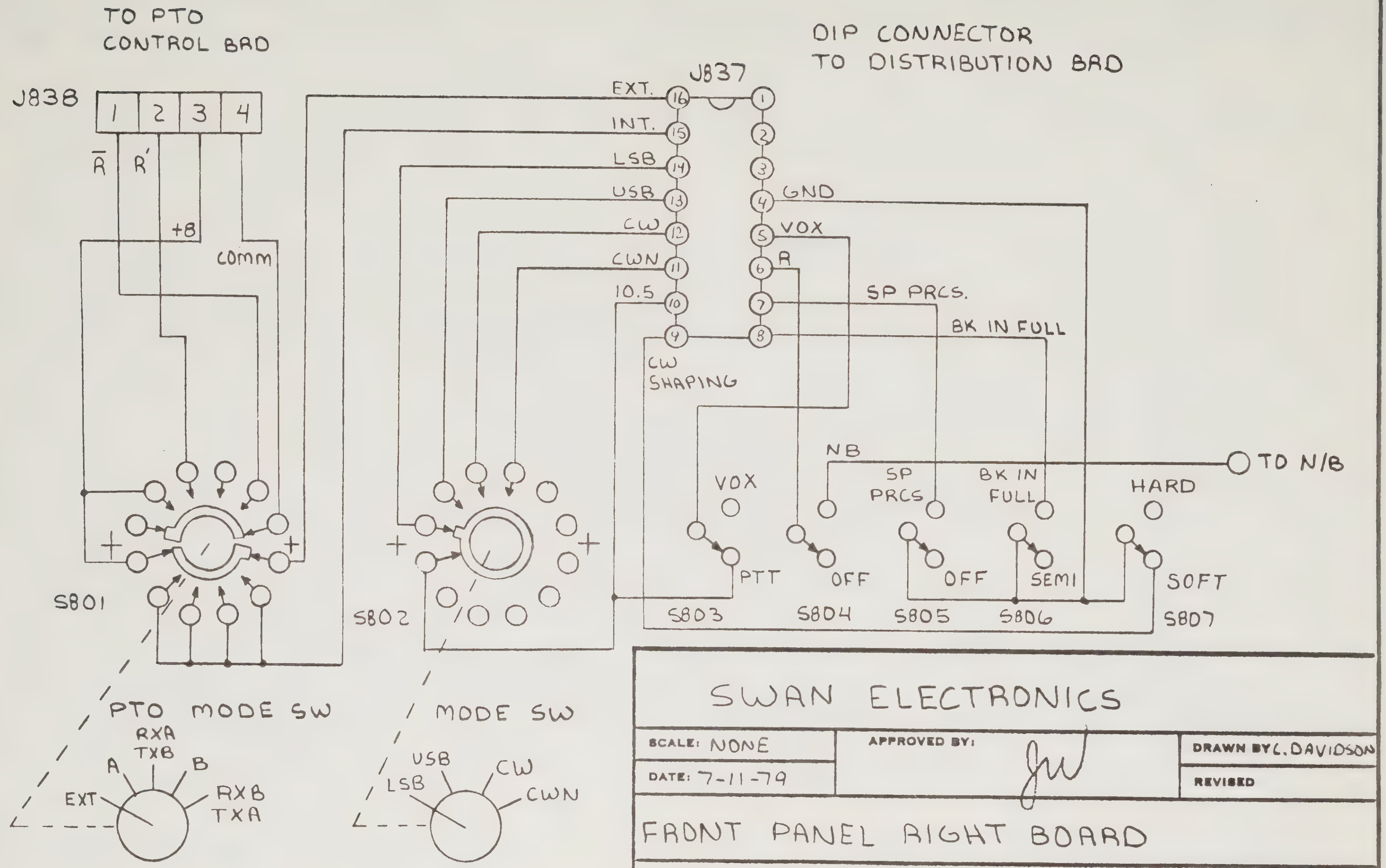
3. J837 & J838 ON FARSIDE.
2. COMPONENT DESIGNATOR SERIES 7800.
1. REF. ELECTRICAL ASSY 102BX-2010
REF. CIRCUIT BOARD 102BX-2110
REF. DRILL DRAWING 102BX-2210
NOTES:

SWAN ELECTRONICS

SCALE: 2 : 1
DATE: 9-12-79
APPROVED BY: JW
DRAWN BY AJG
REVISED

FRONT PANEL RIGHT BD LAYOUT

DRAWING NUMBER 102BX-2310



SWAN ELECTRONICS

SCALE: NONE

DATE: 7-11-79

APPROVED BY: JW

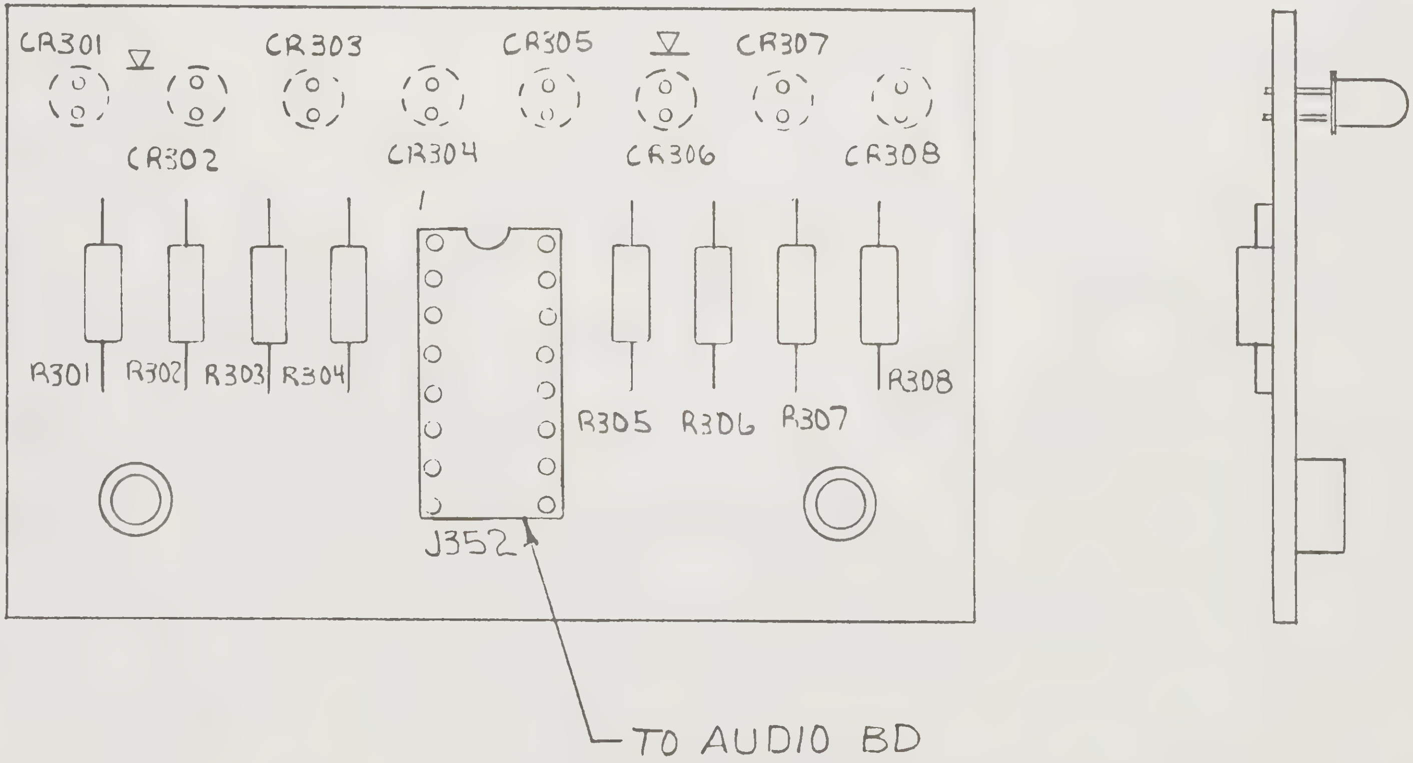
DRAWN BY C. DAVIDSON

REVISED

FRONT PANEL RIGHT BOARD

COMPONENT DESIGNATORS (7800 SERIES)

DRAWING NUMBER 102BX-2410



3. REF. ELECTRICAL ASSY 102BX-2011
   REF. CIRCUIT BOARD 102BX-2111
   REF. DRILL DWG 102BX-2211
2. LEDS MOUNTED FARSIDE.
1. COMPONENT DESIGNATORS 7300 SERIES

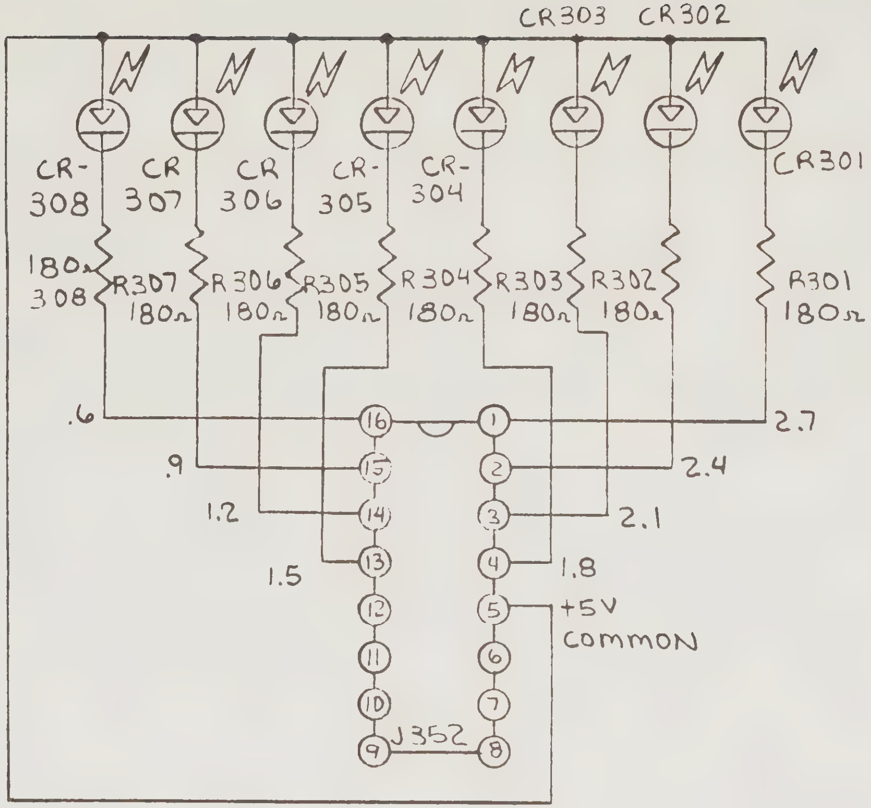
NOTES

SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE:		REVISED

PASSBAND IND. BD LAYOUT
102 BX

DRAWING NUMBER
102BX-2311



DIP CONNECTOR
FROM AUDIO BRD.

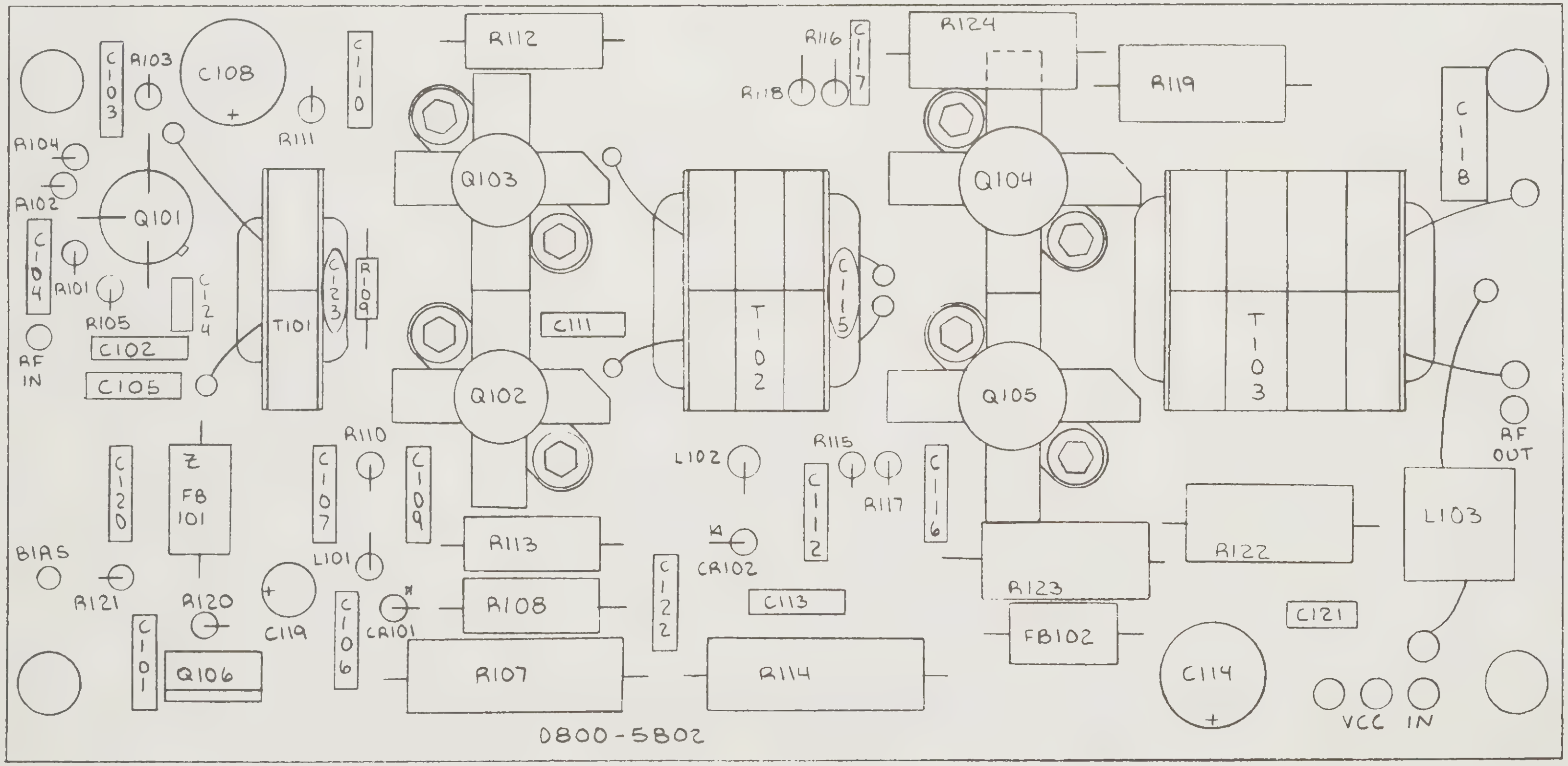
SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: Jw	DRAWN BY C. DAVIDSON
DATE: 7-10-79		REVISED

PASSBAND INDICATOR BOARD SCHEMATIC

DRAWING NUMBER
102B-2411

NOTES:
1. COMPONENT DESIGNATORS 7300 SERIES



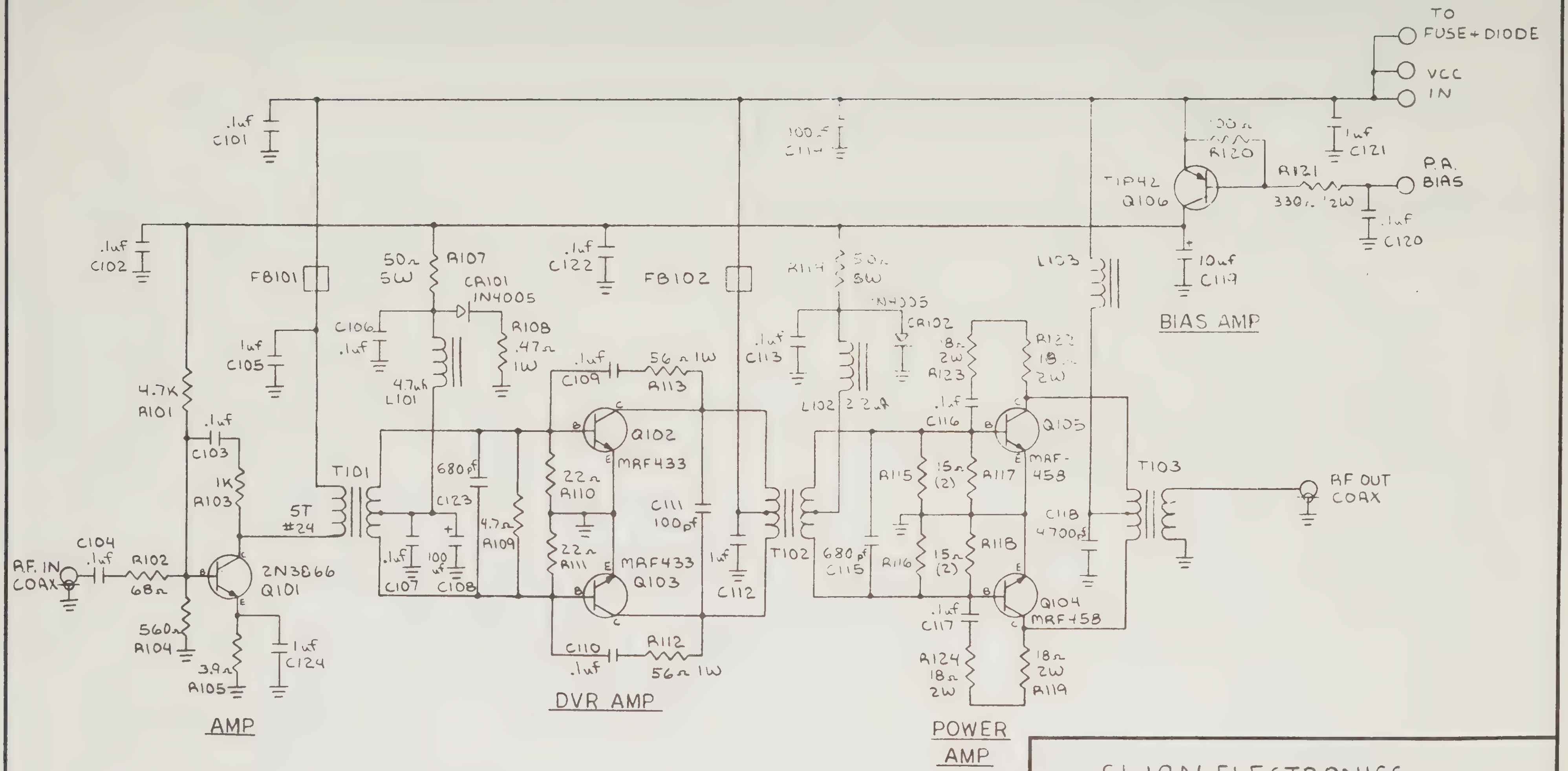
SWAN ELECTRONICS

SCALE: NONE
APPROVED BY:
DRAWN BY: C.DAVIDSON
DATE: 1-2-78
REVISED

POWER AMP BOARD LAYOUT

DRAWING NUMBER
0800-2302

1. COMPONENT DESIGNATORS (6000 SERIES)
NOTES:



ECN			
B	REPLACED ALL .05uf CAPS WITH .1uf CD 5-30-79	6-1-79	JW
LTR	DESCRIPTION	DATE	APP'D

REVISIONS

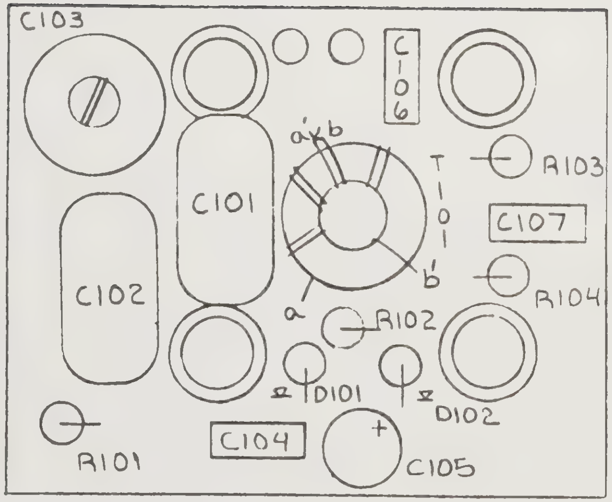
SWAN ELECTRONICS

SCALE: NONE
DATE: 1-2-78
APPROVED BY:
DRAWN BY: C. DAVIDSON
REVISED

POWER AMP BOARD SCHEMATIC

REV B
DRAWING NUMBER 0800-2402B

1. COMPONENT DESIGNATORS (6000 SERIES)
NOTES



SWAN ELECTRONICS

SCALE: NONE
DATE: 1-11-79
APPROVED BY:
DRAWN BY C. DAVIDSON
REVISED

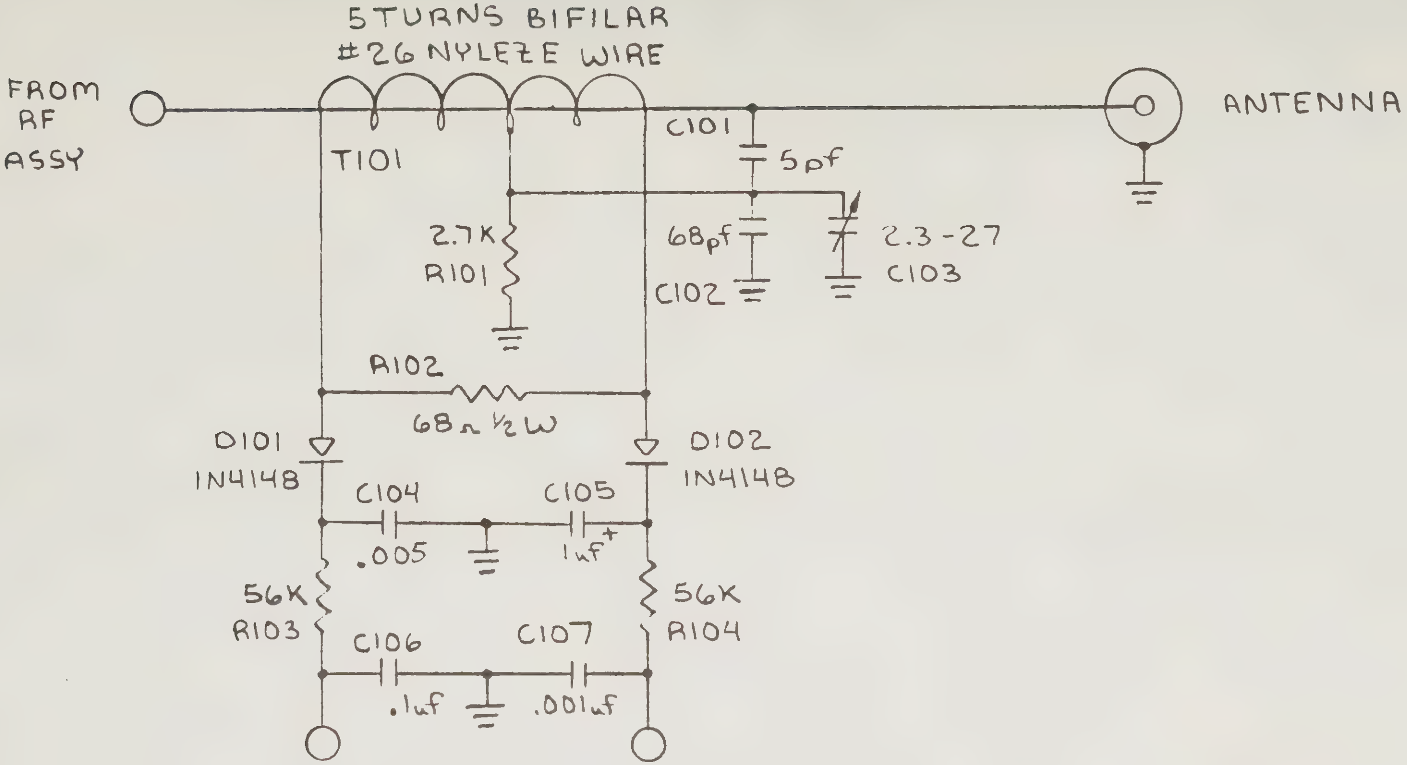
SWR/WATTMETER BRD LAYOUT

DRAWING NUMBER
0800-2303

NOTES:
1. COMPONENT DESIGNATORS (7700 SERIES)

8½X11 PRINTED ON NO. 1000H CLEARPRINT

LTR	DESCRIPTION	DATE	APP'D
A	CHANGED C106 FROM .001 TO .1uF CD 8-24-79	8-24-79	JW



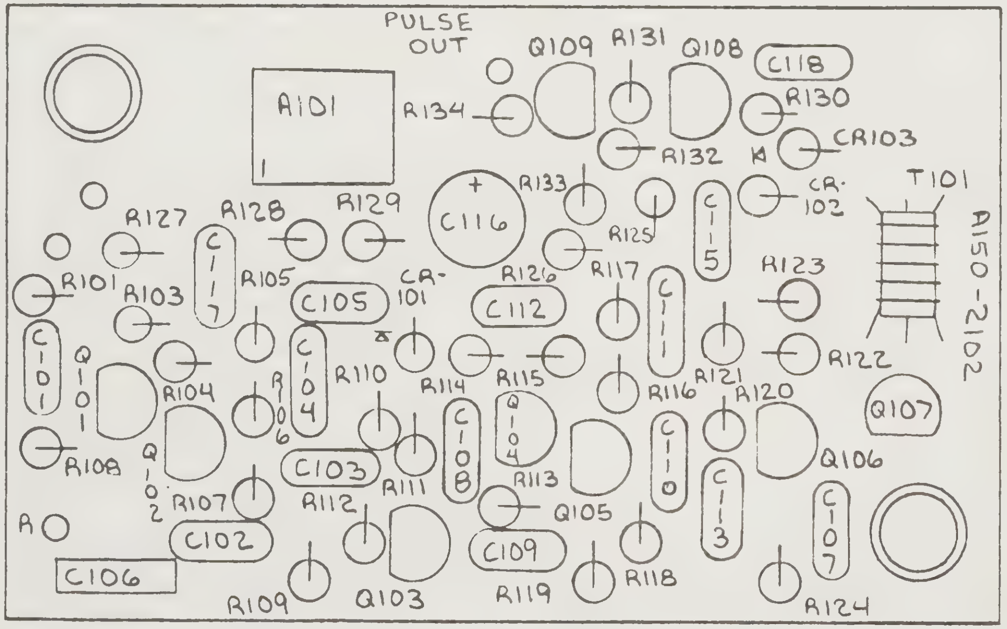
SWAN ELECTRONICS

SCALE: NONE
DATE: 1-11-79
APPROVED BY:
DRAWN BY: C.DAVIDSON
REVISED

SWR/WATTMETER BRD SCHEMATIC

DRAWING NUMBER
0800-2403A

1. COMPONENT DESIGNATORS (7700 SERIES)
NOTES



SWAN ELECTRONICS

SCALE: NONE

DATE: 1-12-79

APPROVED BY: JW

DRAWN BY C.DAVIDSON

REVISED

PROJ.

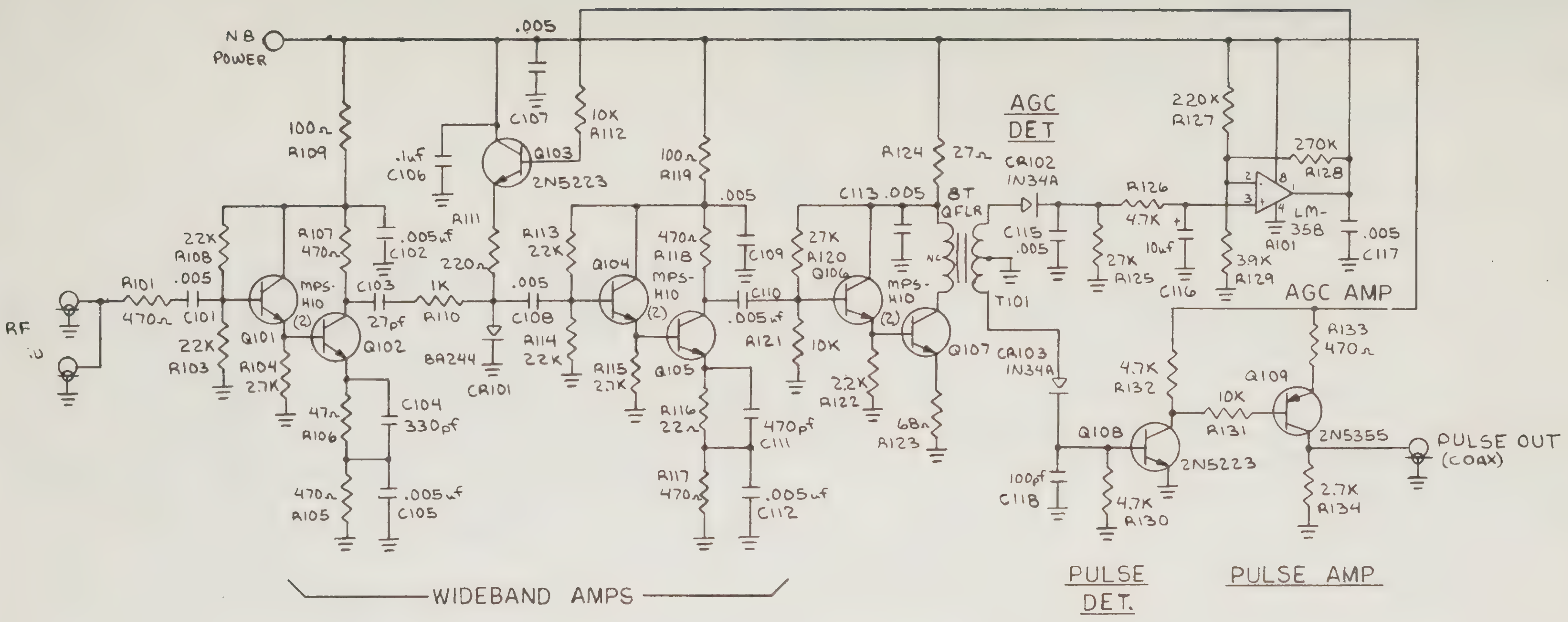
0176

A150 NOISE BLANKER BRD LAYOUT

DRAWING NUMBER
0800-2304

1.COMPONENT DESIGNATORS (7400 SERIES)
NOTES

8½X11 PRINTED ON NO. 1000H CLEARPRINT

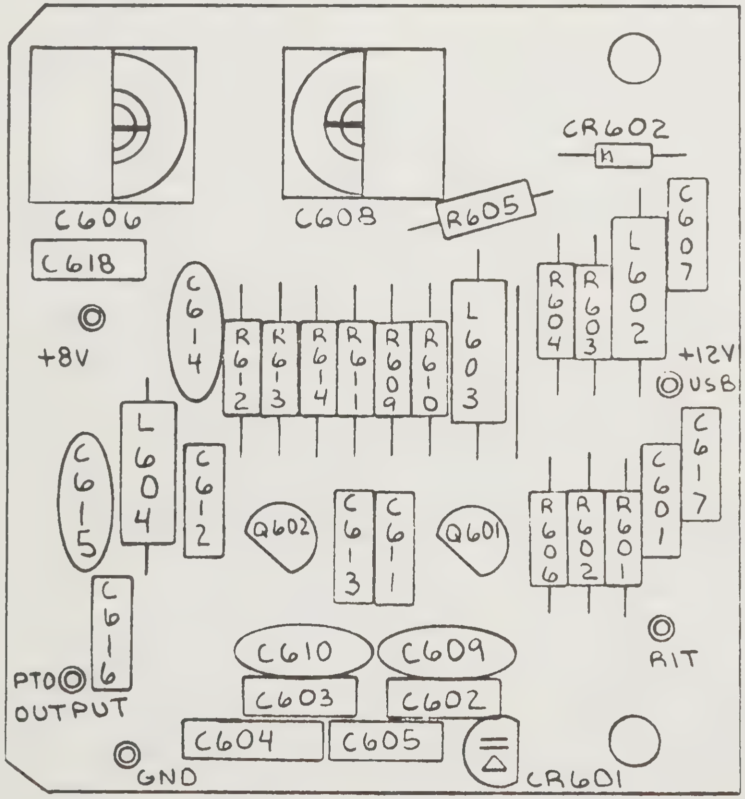


SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY C.DAVID
DATE: 1-12-78		REVISED

PROJ.	A150 NOISE BLANKER BRD SCHEMATIC
0176	DRAWING NUMBER 0800-2404A

1. COMPONENT DESIGNATORS (7400 SERIES)
NOTES



SWAN ELECTRONICS

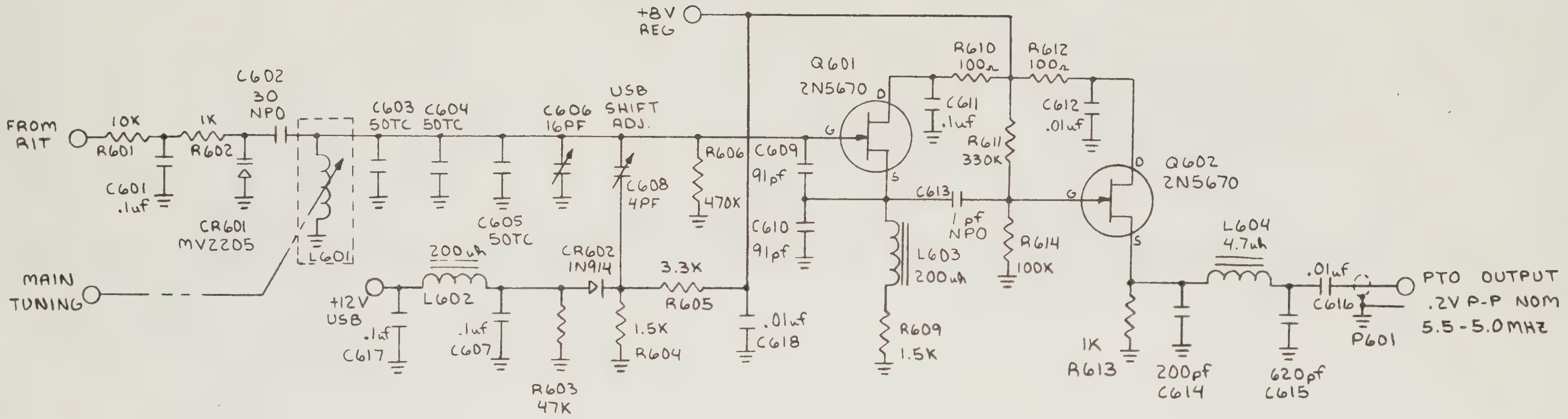
SCALE: NONE
DATE: 9-14-78
APPROVED BY:
DRAWN BY: C DAVIDSON
REVISED

PTO BRD LAYOUT

DRAWING NUMBER
0800-2305

1. COMPONENT DESIGNATORS 7600 SERIES

NOTES



SWAN ELECTRONICS

SCALE: NONE	APPROVED BY:	DRAWN BY C. DAVIDSON
DATE: 9-22-78		REVISED

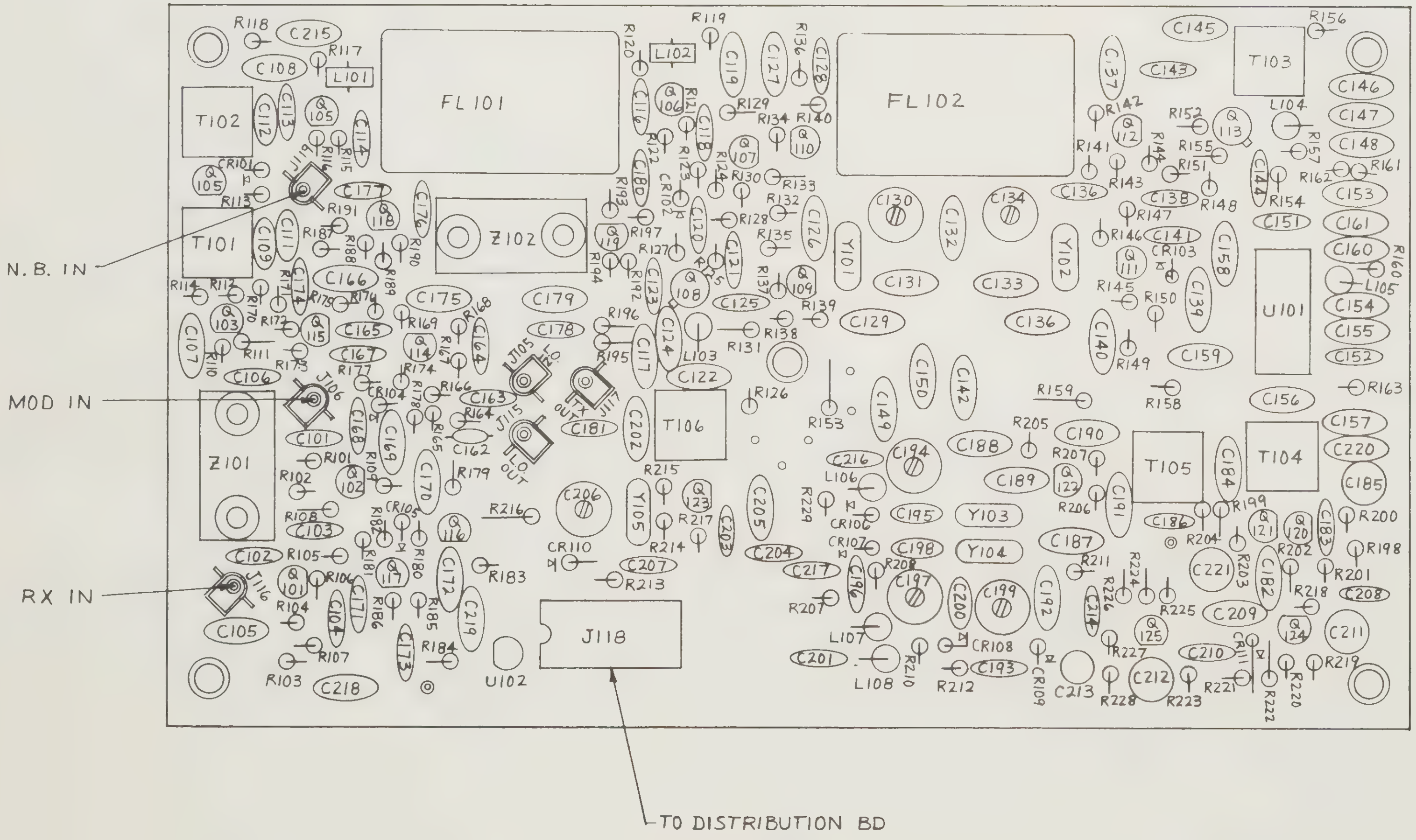
PTO BRD SCHEMATIC (038-188A)

DRAWING NUMBER 0800-2405

1. COMPONENT DESIGNATORS 7600 SERIES

NOTES

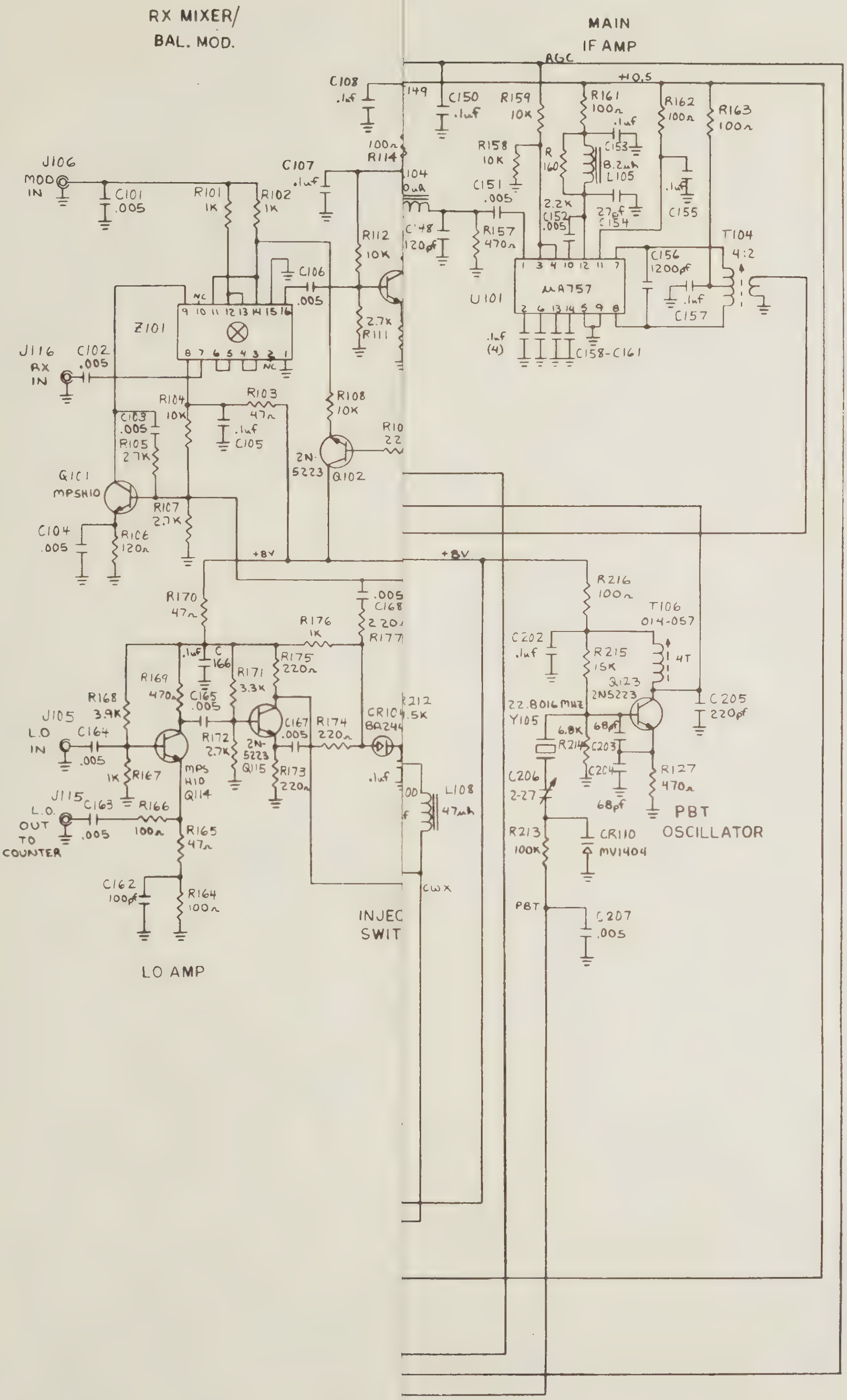
11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS 2000 SERIES.
1. REF. ELECT. ASSY 102BX-2007
REF. CIRCUIT BOARD 102BX-2107
REF. DRILL DWG 102BX-2204

NOTES:

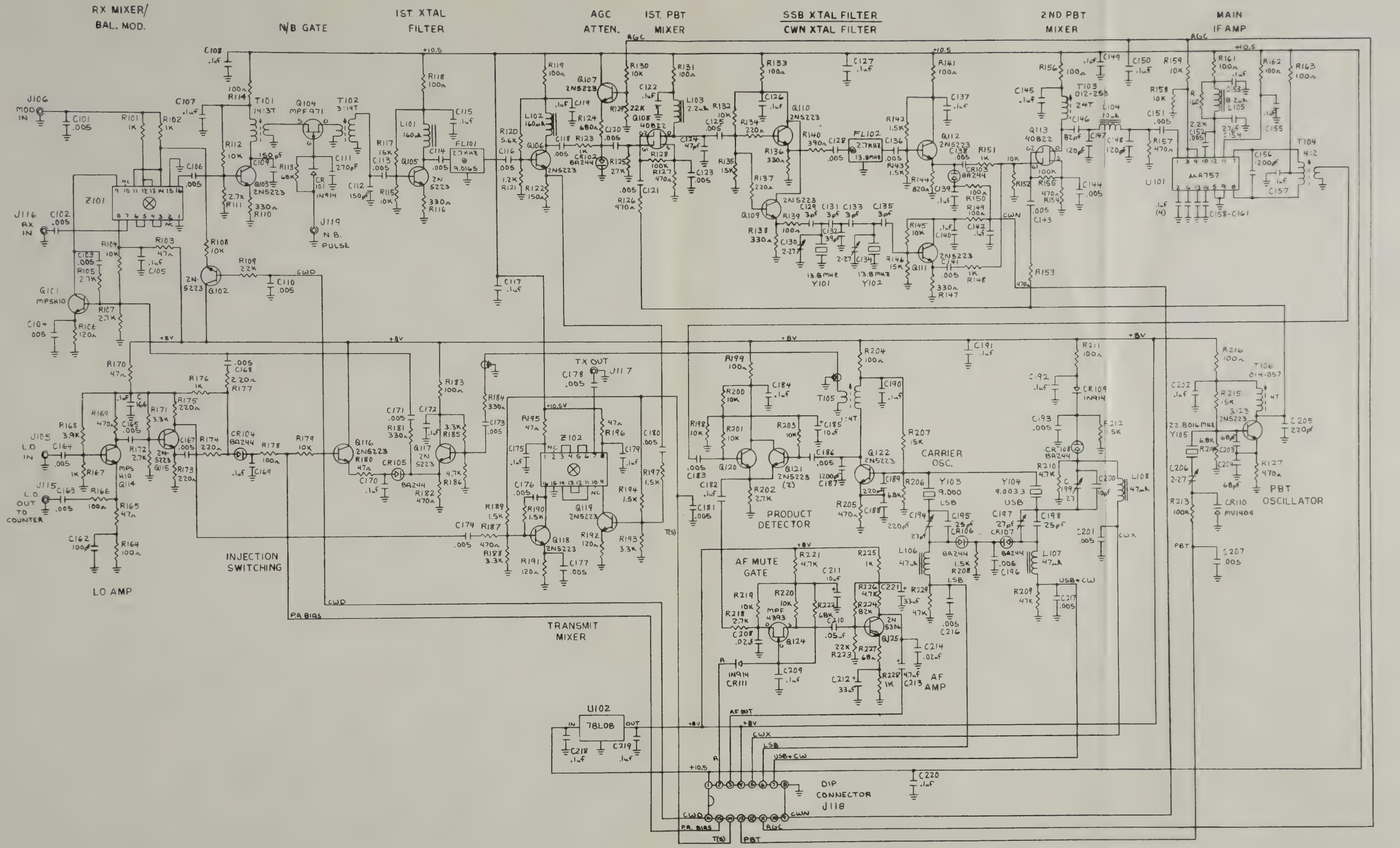
SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 9-24-79		REVISED
EXCITER BOARD LAYOUT		
		DRAWING NUMBER 102BX-2307



SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY: JW	DRAWN BY: C. DRW103
DATE: 8-27-79		REVISED
EXCITER BRD SCHEMATIC		
		DRAWING NUMBER 102B-2407

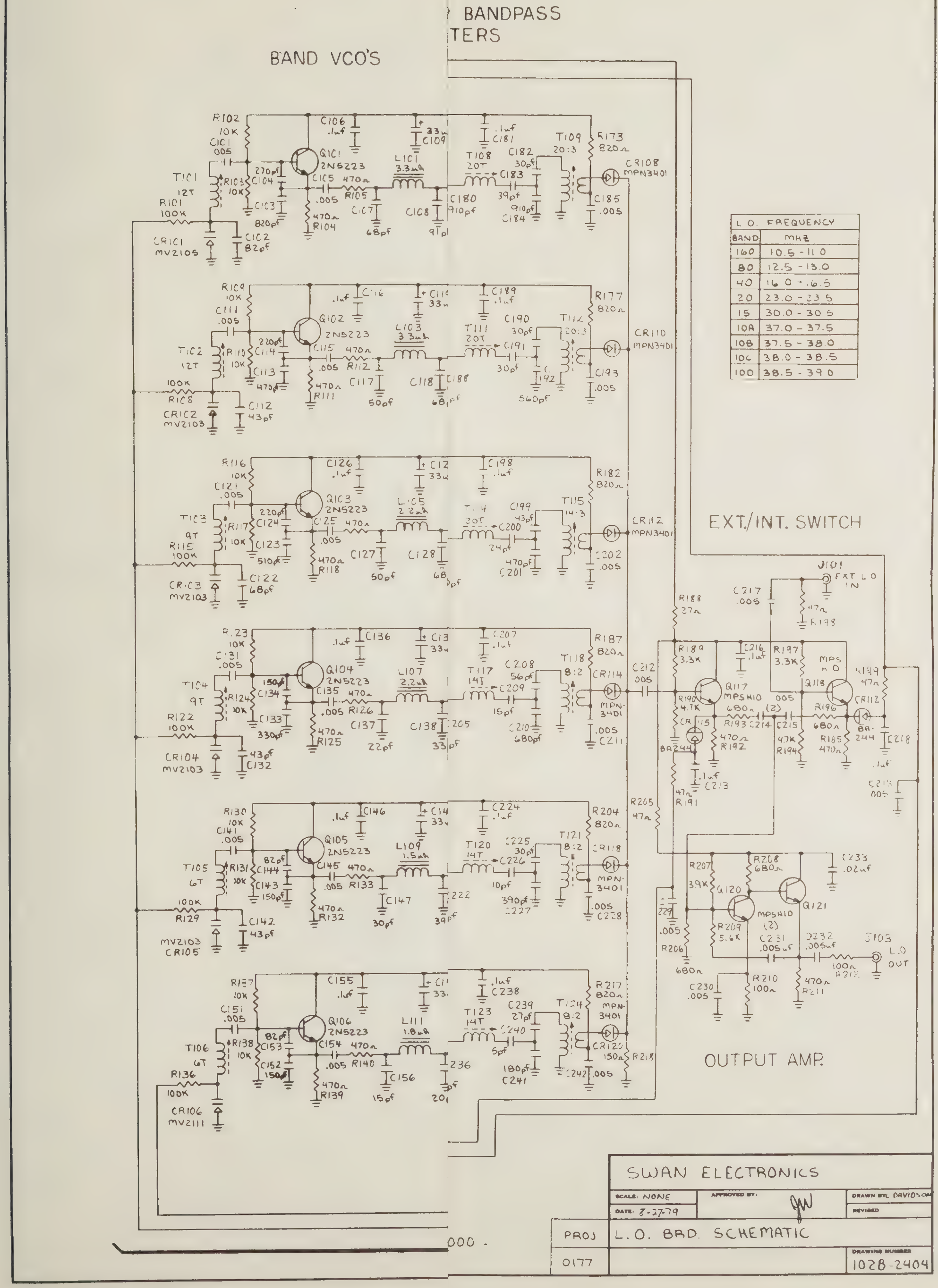
1. COMPONENT DESIGNATORS SERIES

NOTES:



1. COMPONENT DESIGNATORS SERIES 2000.
NOTES:

SWAN ELECTRONICS
SCALE: NONE
DATE 8-27-79
APPROVED BY: JW
EXCITER BRD SCHEMATIC
1028-2407

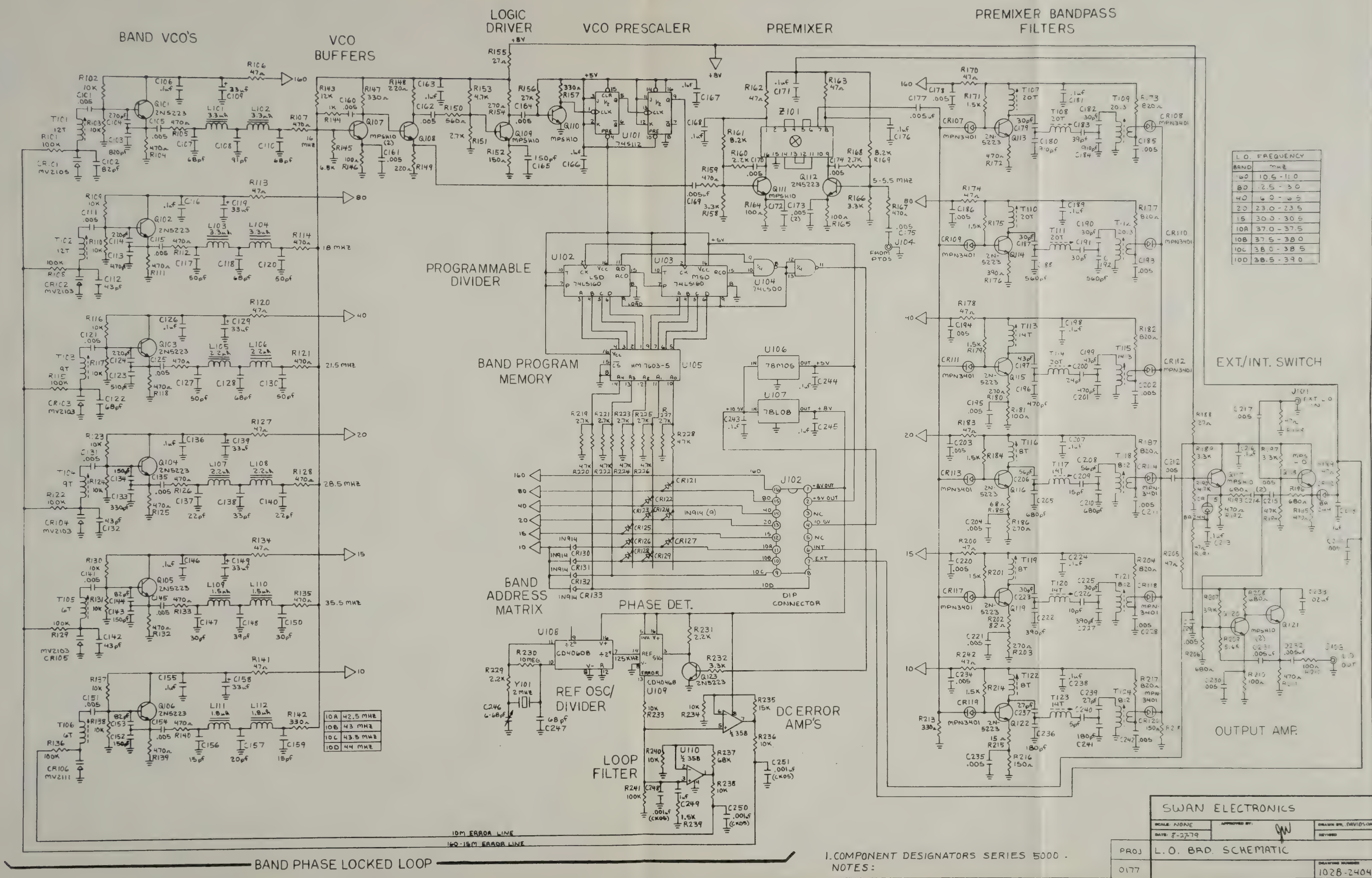


L.O. FREQUENCY	
BAND	MHZ
160	10.5 - 11.0
80	12.5 - 13.0
40	16.0 - 16.5
20	23.0 - 23.5
15	30.0 - 30.5
10A	37.0 - 37.5
10B	37.5 - 38.0
10C	38.0 - 38.5
10D	38.5 - 39.0

SWAN ELECTRONICS

SCALE: NONE
APPROVED BY: JW
DRAWN BY: DAVIDSON
DATE: 8-27-79
REVISED

PROJ 0177
L.O. BRD. SCHEMATIC
DRAWING NUMBER 1028-2404

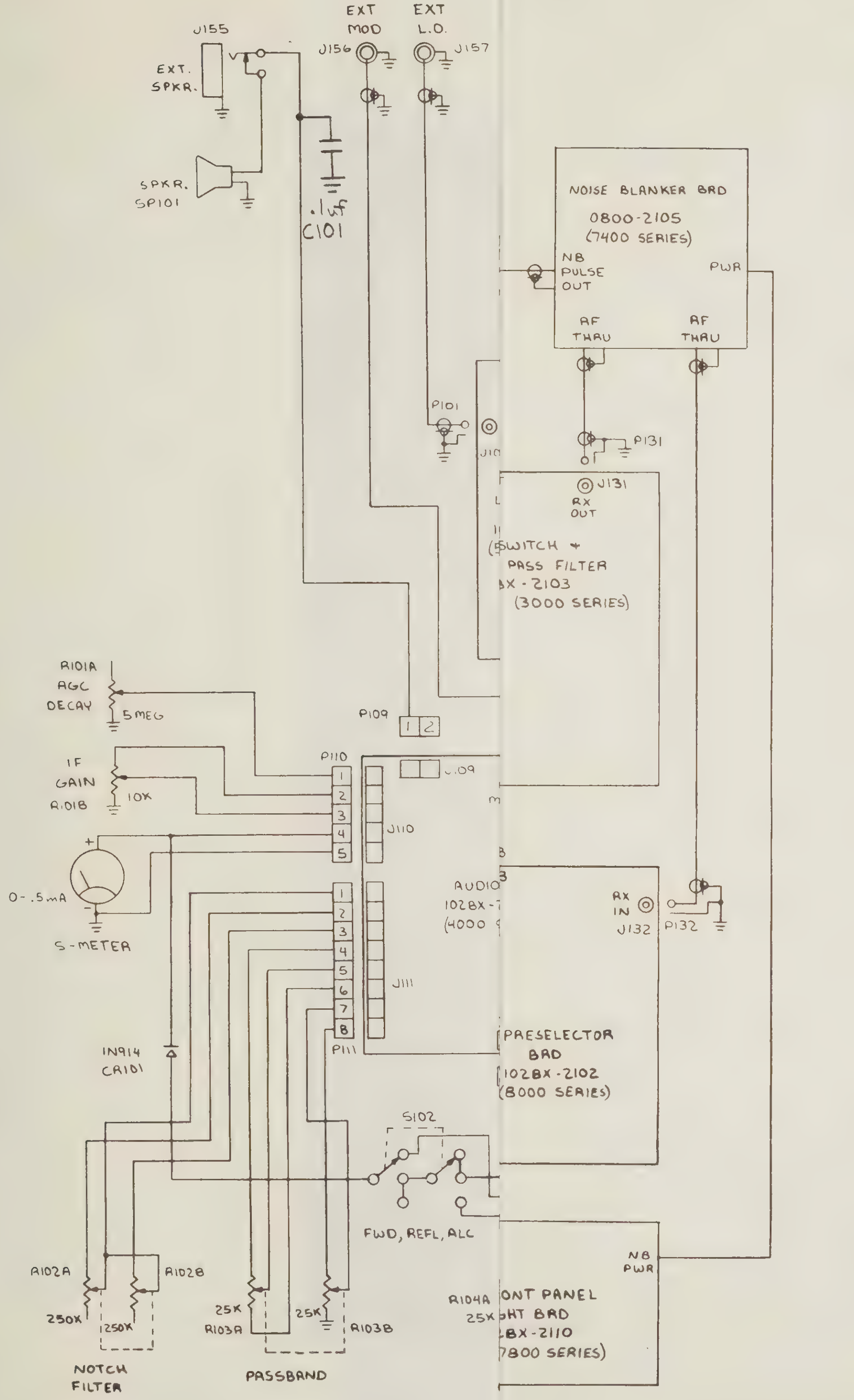


10A	42.5 MHZ
10B	43 MHZ
10C	43.5 MHZ
10D	44 MHZ

L.O. FREQUENCY	
BAND	MHZ
160	10.5 - 11.0
80	12.5 - 13.0
40	6.0 - 6.5
20	23.0 - 23.5
15	30.0 - 30.5
10A	37.0 - 37.5
10B	37.5 - 38.0
10C	38.0 - 38.5
10D	38.5 - 39.0

NOTES:
1. COMPONENT DESIGNATORS SERIES 5000 -

SWAN ELECTRONICS
SCALE: NONE
DATE: 8-27-79
APPROVED BY: JW
DRAWN BY: DAVIDSON
REVISED
PROJ 0177
L.O. BRD. SCHEMATIC
DRAWING NUMBER 1028-2404



NOTES:

1. [N/ ——— N/] DENOTES MULTI CONDUCTOR CABLE ASSY, N = NUMBER OF CONDUCTORS

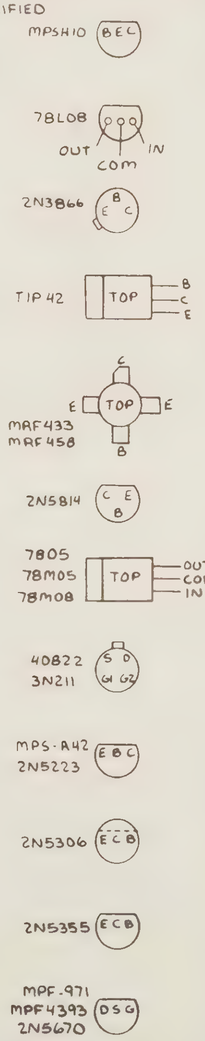
2. [16/ ——— 16/] DENOTES 16 LEAD RIBBON CABLE WITH DIP PLUGS.

3. ALL COMPONENT DESIGNATORS CONSIST OF ONE OR TWO LETTERS FOLLOWED BY FOUR DIGITS.
EXAMPLE: Q6101

Q	6	101
TRANSISTOR	P.A BRD	SEQUENTIAL NUMBER

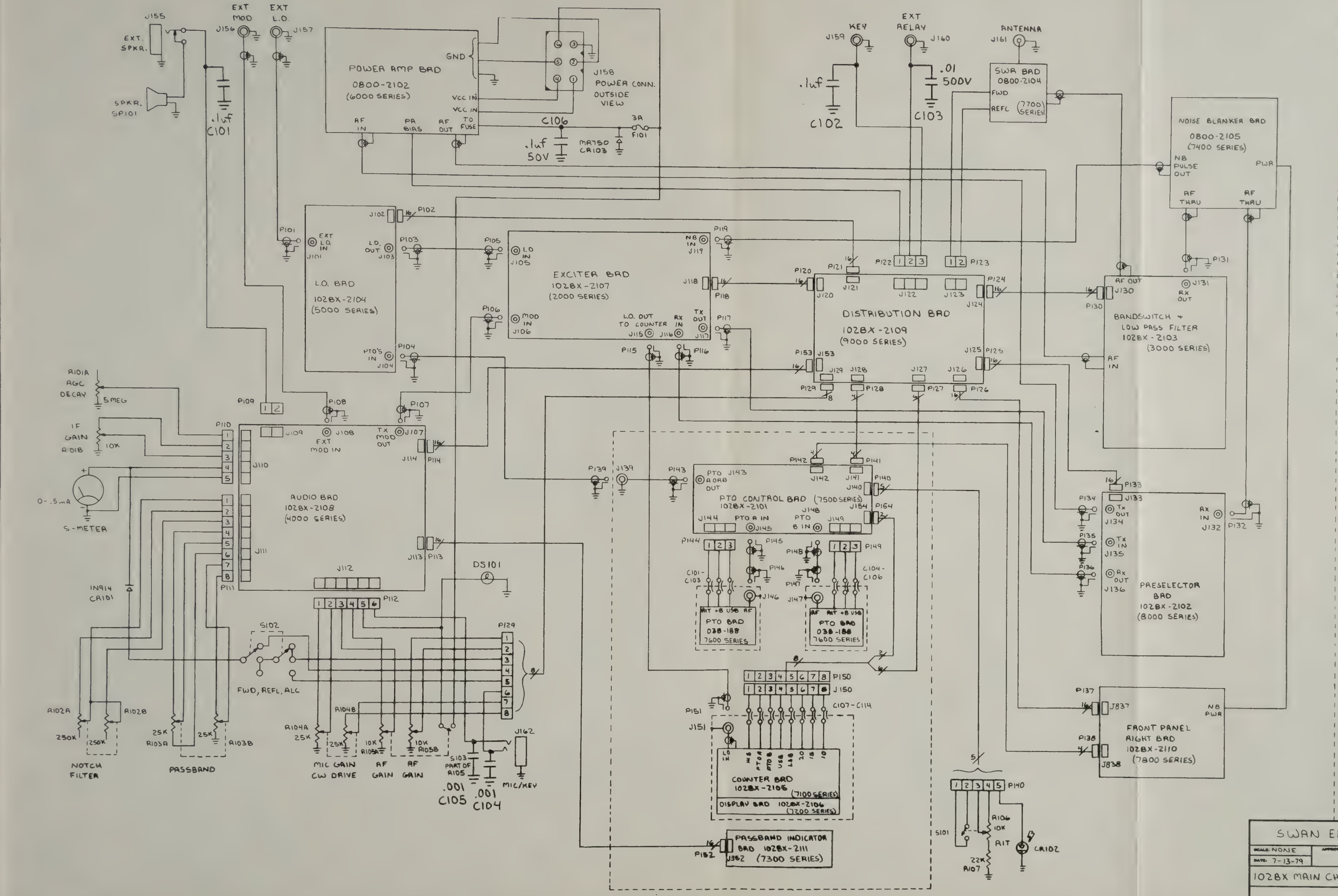
TRANSISTOR DATA

BOTTOM VIEW: UNLESS OTHERWISE SPECIFIED



SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY: C. DAVIDSON
DATE: 7-13-79		REVISED
102BX MAIN CHASSIS + INTERCONNECT		
		DRAWING NUMBER 102BX-2801



NOTES:

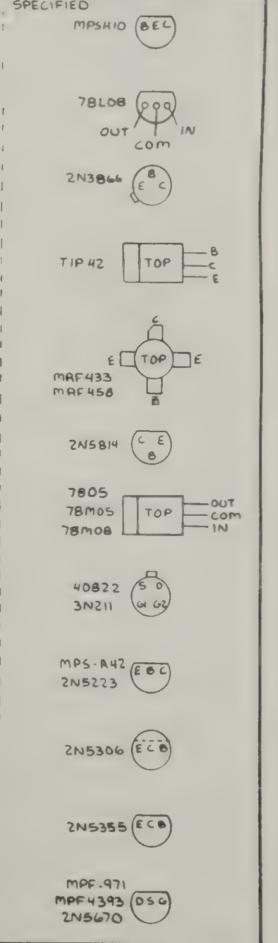
1. N/ N/ DENOTES MULTI CONDUCTOR CABLE ASSY, N = NUMBER OF CONDUCTORS

2. 16/ 16/ DENOTES 16 LEAD RIBBON CABLE WITH DIP PLUGS

3. ALL COMPONENT DESIGNATORS CONSIST OF ONE OR TWO LETTERS FOLLOWED BY FOUR DIGITS.
EXAMPLE: Q6101
Q — TRANSISTOR
6 — P.A BRD
101 — SEQUENTIAL NUMBER

TRANSISTOR DATA

BOTTOM VIEW: UNLESS OTHERWISE SPECIFIED

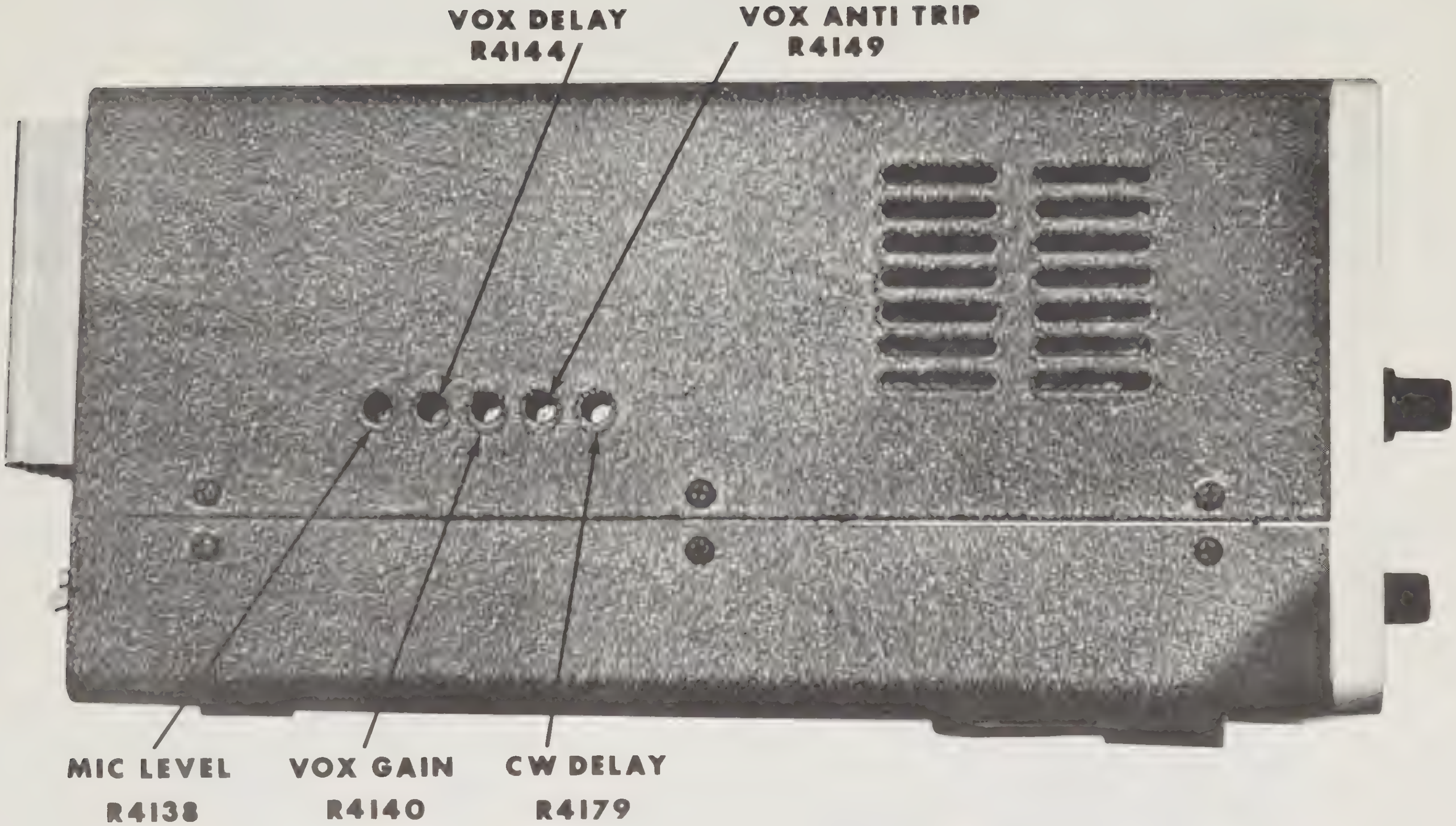


SWAN ELECTRONICS

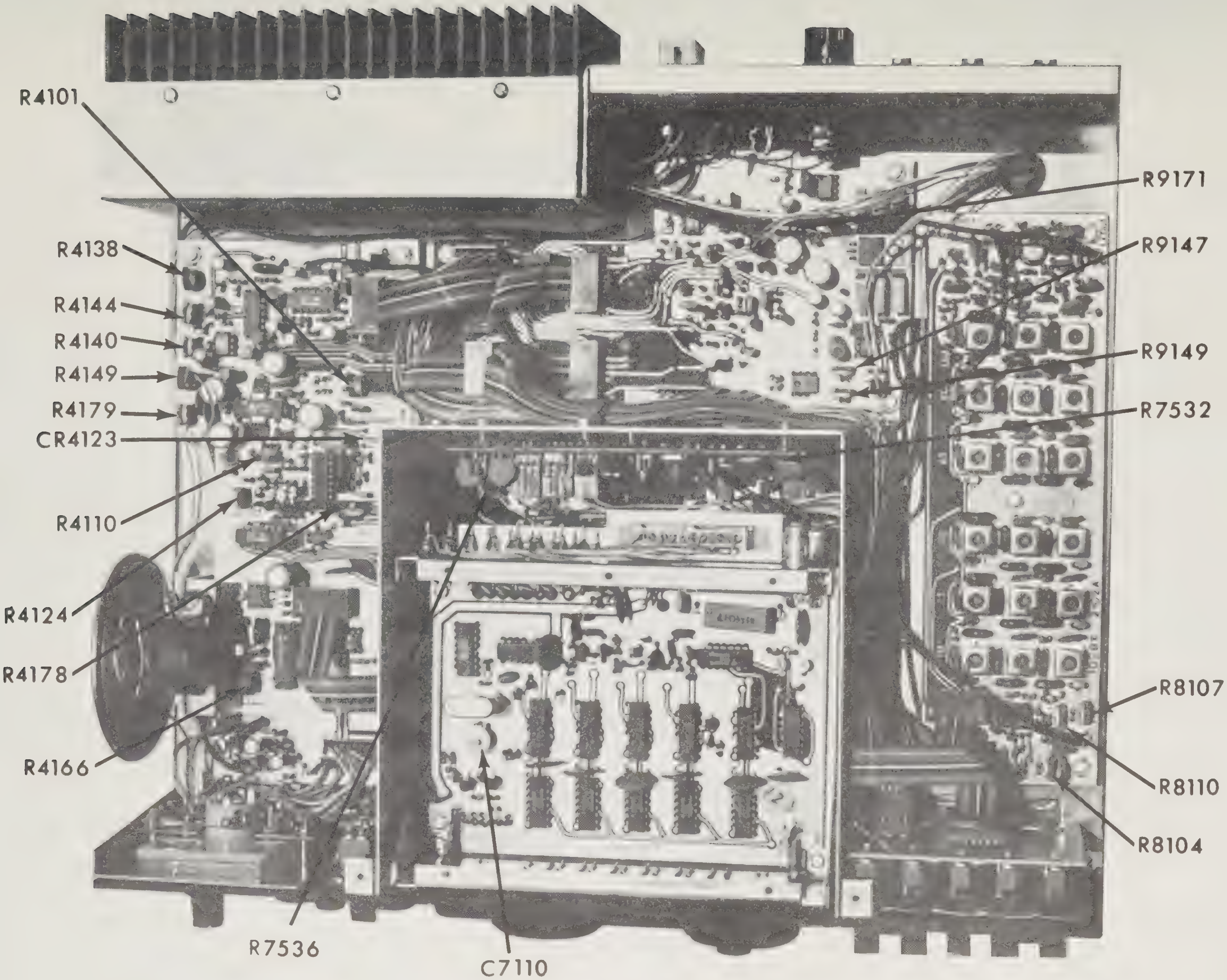
SCALE: NONE
DATE: 7-13-79
APPROVED BY: JW

102BX MAIN CHASSIS + INTERCONNECT

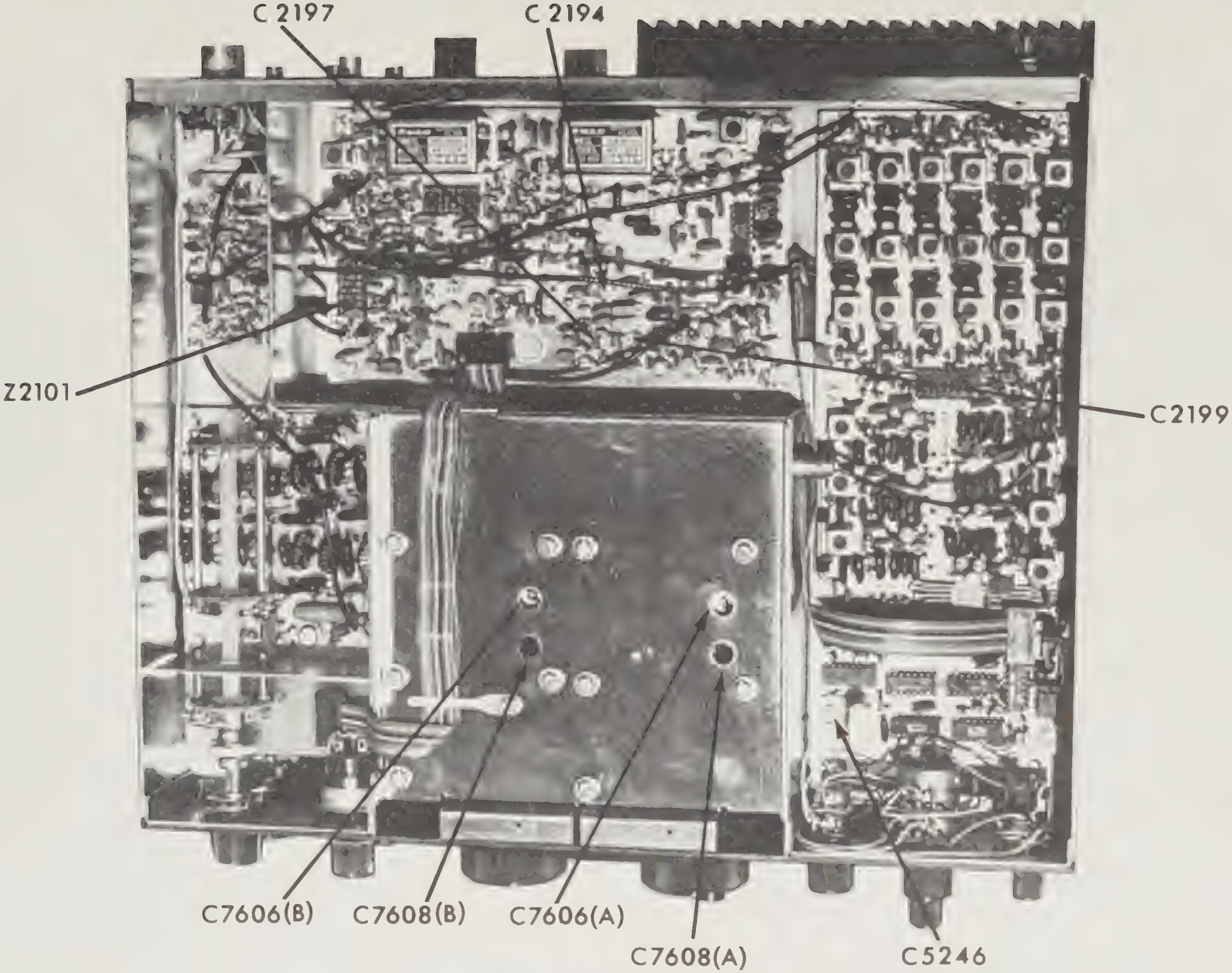
102BX-2801



LEFT SIDE VIEW


R4101
R4138
R4144
R4140
R4149
R4179
CR4123
R4110
R4124
R4178
R4166
R7536
C7110
R9171
R9147
R9149
R7532
R8107
R8110
R8104

TOP VIEW



BOTTOM VIEW

Parts

Semiconductors....................................9-1
Integrated Circuits...............................9-1
Diodes..9-2
Exciter Board (2000 Series).......................9-2
Low Pass Filter/Band Switch Board (3000 Series).....9-3
Audio Board (4000 Series).........................9-4
Local Oscillator Board (5000 Series)..............9-5
Power Amplifier Board (6000 Series)...............9-6
Counter Board (7100 Series).......................9-6
Noise Blanker Board (7400 Series).................9-7
PTO Control Board (7500 Series)...................9-7
PTO Board (7600 Series)...........................9-7
SWR/Wattmeter Board (7700 Series).................9-7
Front Panel Right Board (7800 Series).............9-8
Preselector Board (8000 Series)...................9-8
Distribution Board (9000 Series)..................9-9
Counter Box Assembly..............................9-9
Miscellaneous.....................................9-9

NOTE: Only "parts peculiar" are listed in this section of the manual; that is, only those parts which might otherwise be difficult for the servicing facility to obtain. Common resistors and capacitors are omitted; their value is shown on the schematic diagrams, and the type, size, wattage, etc., can be determined from the part being replaced. Since any service shop can be expected to have a stock of these common parts available, it is unlikely that they would be ordered from Cubic Communications in any event.

Circuit symbols shown on the schematics should be correlated with the 'series' number also shown on the schematic, as only the last three digits of the circuit symbol are used on the diagrams; for instance integrated circuit U101 on the exciter board (2000 series) schematic would be shown in the parts list under circuit symbol U2101; C106 on the low pass filter board (3000 series) would be shown under circuit symbol C3106 in the parts list. The full circuit symbol, the description of the part, and the model and serial number of the radio should be supplied with any parts order to ensure that the correct part is shipped. Semiconductors are listed separately, by manufacturer's type number and Cubic Communications part number only.

SEMICONDUCTORS

Transistors

2N3866	476-015
2N5223	476-077
2N5306	476-024
2N5355	476-016
2N5670	476-039
2N5814	476-017
3N211	476-107
MPF971	476-094
MPF4393	476-060
MPS-A42	476-048
MPS-H10	476-026
MRF433	476-075
MRF458	476-076
TIP-42	476-118
MPS2925	476-074

Integrated Circuits

ADC0804	477-121
CD4016	477-056
CD4046	477-110
CD4060	477-003
CD4081	477-119
LM358	477-072
LM555	477-067
TBA810	477-047
uA741	477-093
uA757	477-069

Integrated Circuits, Continued

74LS00	477-062
74LS10	477-123
74LS90	477-088
74LS112	477-120
74LS160	477-112
74C192	477-192
74LS192	477-082
74196	477-035
78L08	477-074
78M05	477-024
78M08	477-010
7805	677-207
4136	477-111
9374	477-036
HM7603 (L.O. counter)	478-008
HM7603 (BPF indicator)	478-009

Diodes

1N34A	475-008
1N914	475-021
1N4005	475-023
1N5231B	475-061
1N5235A	475-065
1N5236B	475-032
BA244	475-048
MAN3620A	475-087
MPN3401	475-049
MV1404	475-092
MV2103	475-069
MV2105	475-089
MV2111	475-094
MV2205	475-050
MV5174B	475-093

EXCITER BOARD (2000 Series)

Circuit Symbol	Cubic Part Number	Description
C2130	089-028	Capacitor, trimmer, 2-27 pF
C2134	089-028	Capacitor, trimmer, 2-27 pF
C2185	073-052	Capacitor, Electrolytic, 10 uF
C2188	084-084	Capacitor, temp. comp, 220 pF, N2200
C2189	084-084	Capacitor, temp. comp, 220 pF, N2200
C2194	089-028	Capacitor, trimmer, 2-27 pF
C2195	084-111	Capacitor, temp. comp. 25 pF, N470
C2197	089-028	Capacitor, trimmer, 2-27 pF
C2198	084-111	Capacitor, temp. comp. 25 pF, N470
C2199	089-028	Capacitor, trimmer, 2-27 pF
C2206	089-028	Capacitor, trimmer, 2-27 pF

C2211	073-052	Capacitor, electrolytic, 10 uF
C2212	073-085	Capacitor, electrolytic, 33 uF
C2213	073-055	Capacitor, electrolytic, 4.7 uF
C2220	084-060	Capacitor, temp. comp, 10 pF, NPO
C2221	073-085	Capacitor, electrolytic, 33 uF
L2101	012-247	Inductor, 160 uH
L2102	012-247	Inductor, 160 uH
L2103	027-061	RF choke, 2.2 uH
L2104	027-044	RF choke, 10 uH
L2105	027-083	RF choke, 8.2 uH
L2106	027-040	RF choke, 47 uH
L2107	027-040	RF choke, 47 uH
L2108	027-040	RF choke, 47 uH
T2101	014-067	Transformer, RF
T2102	014-067	Transformer, RF
T2103	012-253	Transformer, RF
T2104	014-057	Transformer, RF
T2105	014-058	Transformer, RF
T2106	014-057	Transformer, RF
Y2101	486-102	Crystal, HC18, 13.800 MHz, 20 pF
Y2102	486-102	Crystal, HC18, 13.800 MHz, 20 pF
Y2103	486-080	Crystal, HC18, 9.0000 MHz, 32 pF
Y2104	486-081	Crystal, HC18, 9.0033 MHz, 32 pF
Y2105	486-103	Crystal, HC18, 22.80165 MHz, 10 pF
Z2101	0800-2004	Double balanced mixer
Z2102	0800-2004	Double balanced mixer
FL2101	487-024	Filter, crystal, 9.00165 MHz
FL2102	487-029	Filter, crystal, 13.800 MHz
	341-026	Socket, 16 pin
	344-114	Coax connector, vertical receptical
	344-115	Coax connector, spring socket

LOW PASS FILTER/BAND SWITCH BOARD (3000 Series)

Circuit Symbol	Cubic Part Number	Description
K3101	111-036	Relay, reed, SPST
L3101	027-072	Coil, RF, .35 uH 10 meter
L3102	027-072	Coil, RF, .35 uH, 10 meter
L3103	027-073	Coil, RF, .46 uH, 15 meter
L3104	027-073	Coil, RF, .46 uH, 15 meter
L3105	027-074	Coil, RF, .62 uH, 20 meter
L3106	027-074	Coil, RF, .62 uH, 20 meter
L3107	027-075	Coil, RF, 1.5 uH, 40 meter
L3108	027-075	Coil, RF, 1.5 uH, 40 meter
L3109	027-076	Coil, RF, 2.8 uH, 80 meter
L3110	027-076	Coil, RF, 2.8 uH, 80 meter
L3111	027-079	Coil, RF, 4.5 uH, 160 meter
L3112	027-079	Coil, RF, 4.5 uH, 160 meter

S3101	171-176	Switch, band
S3102	171-176	Switch, band
S3103	171-176	Switch, band
	341-026	Socket, 16 pin
	344-114	Coax connector, vertical receptical
	344-115	Coax connector, spring socket

AUDIO BOARD (4000 Series)

Circuit Symbol	Cubic Part Number	Description
C4102	069-024	Capacitor, tantalum, 68 uF
C4103	073-085	Capacitor, electrolytic, 33 uF
C4105	070-001	Capacitor, stable ceramic, .0047 uF
C4106	070-001	Capacitor, stable ceramic, .0047 uF
C4107	073-055	Capacitor, electrolytic, 4.7 uF
C4108	073-064	Capacitor, electrolytic, 100 uF
C4109	073-064	Capacitor, electrolytic, 100 uF
C4112	073-064	Capacitor, electrolytic, 100 uF
C4115	073-064	Capacitor, electrolytic, 100 uF
C4116	070-012	Capacitor, stable ceramic, 1 uF
C4121	073-085	Capacitor, electrolytic, 33 uF
C4122	073-051	Capacitor, electrolytic, 1 uF
C4123	073-085	Capacitor, electrolytic, 33 uF
C4125	073-052	Capacitor, electrolytic, 10 uF
C4126	073-085	Capacitor, electrolytic, 33 uF
C4128	073-051	Capacitor, electrolytic, 1 uF
C4130	073-052	Capacitor, electrolytic, 10 uF
C4131	073-051	Capacitor, electrolytic, 1 uF
C4132	073-051	Capacitor, electrolytic, 1 uF
C4133	073-051	Capacitor, electrolytic, 1 uF
C4134	073-051	Capacitor, electrolytic, 1 uF
C4136	073-051	Capacitor, electrolytic, 1 uF
C4139	073-051	Capacitor, electrolytic, 1 uF
C4140	073-051	Capacitor, electrolytic, 1 uF
C4141	073-088	Capacitor, electrolytic, 220 uF
C4143	070-001	Capacitor, stable ceramic, .0047 uF
C4144	088-137	Capacitor, silver mica, 1800 pF, 1%, DM19
C4145	088-137	Capacitor, silver mica, 1800 pF, 1%, DM19
R4101	052-119	Trimpot, 10 kΩ
R4110	052-119	Trimpot, 10 kΩ
R4124	052-131	Trimpot, 1 kΩ
R4138	052-119	Trimpot, 10 kΩ
R4140	052-133	Trimpot, 1 MΩ
R4144	052-133	Trimpot, 1 MΩ
R4149	052-148	Trimpot, 100 Ω
R4166	052-151	Trimpot, 50 kΩ
R4174	040-005	Resistor, 47 kΩ, $\frac{1}{4}$ watt, 1%
R4175	040-004	Resistor, 10 kΩ, $\frac{1}{4}$ watt, 1%
R4176	040-006	Resistor, 5.6 kΩ, $\frac{1}{4}$ watt, 1%
R4178	052-151	Trimpot, 50 kΩ
R4179	052-133	Trimpot, 1 MΩ

341-026	Socket, 16 pin
344-114	Coax connector, vertical receptical
344-115	Coax connector, spring socket

LOCAL OSCILLATOR BOARD (5000 Series)

Circuit Symbol	Cubic Part Number	Description
C5109	073-085	Capacitor, electrolytic, 33 uF
C5119	073-085	Capacitor, electrolytic, 33 uF
C5129	073-085	Capacitor, electrolytic, 33 uF
C5139	073-085	Capacitor, electrolytic, 33 uF
C5149	073-085	Capacitor, electrolyitc, 33 uF
C5158	073-085	Capacitor, electrolytic, 33 uF
C5246	089-030	Capacitor, trimmer, 3.5-65 pF
C5248	070-002	Capacitor, stable ceramic, .001 uF
C5249	070-012	Capacitor, stable ceramic, 1 uF
C5250	070-002	Capacitor, stable ceramic, .001 uF
C5251	070-002	Capacitor, stable ceramic, .001 uF
L5101	027-069	RF Choke, 3.3 uH
L5102	027-069	RF Choke, 3.3 uH
L5103	027-069	RF Choke, 3.3 uH
L5104	027-069	RF Choke, 3.3 uH
L5105	027-061	RF Choke, 2.2 uH
L5106	027-061	RF Choke, 2.2 uH
L5107	027-061	RF Choke, 2.2 uH
L5108	027-061	RF Choke, 2.2 uH
L5109	027-059	RF Choke, 1.5 uH
L5110	027-059	RF Choke, 1.5 uH
L5111	027-060	RF Choke, 1.8 uH
L5112	027-060	RF Choke, 1.8 uH
T5101	012-254	Transformer, RF
T5102	012-254	Transformer, RF
T5103	012-255	Transformer, RF
T5106	012-255	Transformer, RF
T5107	014-068	Transformer, RF
T5108	014-068	Transformer, RF
T5109	014-068	Transformer, RF
T5110	014-068	Transformer, RF
T5111	014-068	Transformer, RF
T5112	014-068	Transformer, RF
T5113	014-067	Transformer, RF
T5114	014-068	Transformer, RF
T5115	014-067	Transformer, RF
T5116	014-066	Transformer, RF
T5117	014-067	Transformer, RF
T5118	014-066	Transformer, RF
T5119	014-066	Transformer, RF
T5120	014-067	Transformer, RF
T5121	014-066	Transformer, RF
T5122	014-066	Transformer, RF

T5123	014-067	Transformer, RF
T5124	014-066	Transformer, RF
Y5101	486-100	Crystal, HC33, 2.000 MHz, ± .005%
Z5101	0800-039	Double balanced mixer
	341-026	Socket, 16 pin
	344-114	Coax connector, vertical receptical
	344-115	Coax connector, spring socket

POWER AMPLIFIER BOARD (6000 Series)

Circuit Symbol	Cubic Part Number	Description
C6105	070-012	Capacitor, stable ceramic, 1 uF
C6108	073-064	Capacitor, electrolytic, 100 uF
C6112	070-012	Capacitor, stable ceramic, 1 uF
C6114	073-064	Capacitor, electrolytic, 100 uF
C6119	073-060	Capacitor, electrolytic, 10 uF
C6121	070-012	Capacitor, stable ceramic, 1 uF
C6124	070-012	Capacitor, stable ceramic, 1 uF
FB6101	027-055	Ferrite bead
FB6102	027-055	Ferrite bead
L6101	027-050	RF choke, 4.7 uH
L6102	027-061	RF choke, 2.2 uH
L6103	027-071	RF choke
R6107	049-060	Resistor, 50Ω 5 W, 10%
R6108	049-017	Resistor, 0.47Ω, 2 W, 5%
R6114	049-060	Resistor, 50Ω 5 W, 10%
R6119	048-180	Resistor, 18Ω, 2 W, 5%
R6122	048-180	Resistor, 18Ω, 2 W, 5%
R6123	048-180	Resistor, 18Ω, 2 W, 5%
R6124	048-180	Resistor, 18Ω, 2 W, 5%
T6101	273-014	Transformer, RF, interstage
T6102	273-013	Transformer, RF, interstage
T6103	273-006B	Transformer, RF, output
	116-002	Heatsink
	116-019	Heatsink, power amplifier

COUNTER BOARD (7100 Series)

Circuit Symbol	Cubic Part Number	Description
C7102	070-012	Capacitor, stable ceramic, 1 uF
C7110	089-030	Capacitor, trimmer, 3.5 to 65 pF
C7111	084-116	Capacitor, temp. comp, 68 pF, N5200
R7101	049-068	Resistor, wirewound, 4.7Ω, 5 W.
Y7101	486-101	Crystal, 1.31072 MHz, HC33, 20 pF parallel

NOISE BLANKER BOARD (7400 Series)

Circuit Symbol	Cubic Part Number	Description
C7416	073-060	Capacitor, electrolytic, 10 uF
T7401	014-062	Transformer, RF

PTO CONTROL BOARD (7500 Series)

Circuit Symbol	Cubic Part Number	Description
L7501	027-078	RF Choke, 24 uH
L7502	027-078	RF Choke, 24 uH
L503	027-044	RF Choke, 10 uH
R7532	052-119	Trimpot, 10 kΩ
R7536	052-119	Trimpot, 10 kΩ
	344-114	Coax connector, vertical receptical
	344-115	Coax connector, spring socket
	344-117	Wafer connector, 4 pin
	344-147	Wafer connector, 3 pin
	344-148	Wafer connector, 5 pin

PTO BOARD (7600 Series)

Circuit Symbol	Cubic Part Number	Description
C7602	084-011	Capacitor, temp. comp, 30 pF, NPO
C7603	084-057	Capacitor, temp. comp, 50 pF, N3300
C7604	084-057	Capacitor, temp. comp, 50 pF, N3300
C7605	084-088	Capacitor, temp. comp, 50 pF, N75
C7606	075-005	Capacitor, trimmer, 1.8 to 13 pF
C7608	075-017	Capacitor, trimmer, 1.2 to 4.2 pF
C7613	084-093	Capacitor, temp. comp., 1 pF NPO
L7602	027-049	RF choke, 200 uH
L7603	027-049	RF choke, 200 uH
L7604	027-050	RF choke, 4.7 uH

SWR/WATTMETER BOARD (7700 Series)

Circuit Symbol	Cubic Part Number	Description
C7703	089-028	Capacitor, trimmer, 4-30 pF
C7705	073-051	Capacitor, electrolytic, 1 uF
T7701	012-256	Transformer, RF
	0800-5303	Coax receptical, modified S0239

FRONT PANEL RIGHT BOARD (7800 Series)

Circuit Symbol	Cubic Part Number	Description
S7801	171-188	Switch, PTO Mode
S7802	171-189	Switch, Mode
S7803	172-035	Switch, VOX/PTT
S7804	172-035	Switch, Noise Blanker
S7805	172-035	Switch, Speech Processor
S7806	172-035	Switch, Breakin
S7807	172-035	Switch, Hard/Soft Keying
	344-117	Wafer connector, 4 circuit
	341-026	Socket, 16 pin

PRESELECTOR BOARD (8000 Series)

Circuit Symbol	Cubic Part Number	Description
L8101	027-078	RF Choke, 24 uH
L8102	027-078	RF Choke, 24 uH
L8103	012-247	RF Choke, 160 uH
L8104	012-257	Transformer, RF
R8104	052-133	Trimpot, 1 MΩ
R8107	052-133	Trimpot, 1 MΩ
R8110	052-132	Trimpot, 100 kΩ
T8101	014-070	Transformer, RF
T8102	014-070	Transformer, RF
T8103	014-070	Transformer, RF
T8104	014-069	Transformer, RF
T8105	014-069	Transformer, RF
T8106	014-069	Transformer, RF
T8107	014-068	Transformer, RF
T8108	014-069	Transformer, RF
T8109	014-068	Transformer, RF
T8110	014-067	Transformer, RF
T8111	014-068	Transformer, RF
T8112	014-067	Transformer, RF
T8113	014-066	Transformer, RF
T8114	014-067	Transformer, RF
T8115	014-066	Transformer, RF
T8116	014-066	Transformer, RF
T8117	014-067	Transformer, RF
T8118	014-066	Transformer, RF
	344-114	Coax connector, vertical receptacle
	344-115	Coax connector, spring socket
	341-026	Socket, 16 contact

DISTRIBUTION BOARD (9000 Series)

Circuit Symbol	Cubic Part Number	Description
C9101	070-012	Capacitor, stable ceramic, 1 uF
C9102	070-012	Capacitor, stable ceramic, 1 uF
C9103	070-013	Capacitor, stable ceramic, 0.1 uF
C9104	070-012	Capacitor, stable ceramic, 1 uF
C9105	070-020	Capacitor, stable ceramic, 0.33 uF
C9106	069-017	Capacitor, tantalum, 2.2 uF
C9108	070-012	Capacitor, stable ceramic, 1 uF
C9109	073-088	Capacitor, electrolytic, 220 uF
C9111	073-088	Capacitor, electrolytic, 220 uF
C9114	073-085	Capacitor, electrolytic, 33 uF
C9115	073-085	Capacitor, electrolytic, 33 uF
C9116	073-085	Capacitor, electrolytic, 33 uF
C9120	073-055	Capacitor, electrolytic, 4.7 uF
R9141	052-119	Trimpot, 10 kΩ
R9147	052-133	Trimpot, 1 MΩ
R9149	052-133	Trimpot, 1 MΩ
	116-020	Heat sink for TIP42
	311-014	Insulating washer
	341-026	Socket, 16 pin
	344-117	Wafer connection, 4 pin
	344-136	Wafer connection, 8 pin
	344-150	Wafer connection, 2 pin
	344-147	Wafer connection, 3 pin
	344-149	Wafer connection, 6 pin

COUNTER BOX ASSEMBLY

Cubic Part Number	Description
077-001	Capacitor, feed through, 1000 pF
344-146	Connector, 3 circuit
344-151	Receptacle, 8 circuit
344-152	Plug, 8 circuit
344-153	Crimp terminal for plug
344-154	Crimp terminal for socket

MISCELLANEOUS

Cubic Part Number	Description
222-001	Rubber feet, small
223-002	Tilt stand, 10"
102B-1310	Cabinet top
102B-1311	Cabinet bottom
0800-5403	Bezel
150-032	D.C. Power cord
222-002	Rubber feet, large

045-223F	Potentiometer, RIT, with switch
052-112	Potentiometer, AF/RF Gain, with switch
052-152	Potentiometer, MIC Gain and Passband Tuning
052-153	Potentiometer, IF and AGC
052-154	Potentiometer, tandem, Notch Filter
112-038	Meter
152-008	Speaker
171-190	Bandswitch detent
172-036	Switch, rocker, level handle
211-043	Knob, 3/4" with insert
211-044	Wing knob, 3/4", with insert
211-072	Knob, dual, 4mm and 6mm shafts
211-073	Knob, 3/4", with insert; .187 dia shaft
211-047	Knob, dual, with bar
211-080	Knob, with spinner
344-082	Receptacle, 6 circuit, power
344-083	Pins for above
344-118	Connector, 4 pin, housing only
344-135	Connector, 8 pin, housing only
344-146	Connector, 3 pin, housing only
344-155	Connector, 2 pin, housing only
344-156	Connector, 5 pin, housing only
344-157	Connector, 6 pin, housing only
344-119	Crimp terminal for above
455-123	Escutcheon
455-124	Filter for digital readout
471-007	Lamp, 12 volt, miniature
680-004	Ribbon cable assembly, 3 3/4"
680-005	Ribbon cable assembly, 9"
680-006	Ribbon cable assembly, 18 1/8"
680-007	Ribbon cable assembly, 11 1/2"
680-009	Ribbon cable assembly, 15"
680-010	Ribbon cable assembly, 16 1/8"
102B-3203	PTO Assembly, complete

OPERATING MANUAL

MODEL PSU-6 POWER SUPPLY

1. The Swan PSU-6 fixed station power supply-speaker provides a regulated D.C. voltage for operation of all Swan solid state transceivers, but is intended particularly to match the Swan ASTRO-102BX Transceiver.

2. The PSU-6 is designed to operate from nominal 117 volts A.C., 50-60 Hz or 230 volts A.C. 50-60 Hz. All assemblies are factory wired for 117 v.a.c. operation, however, they may easily be re-wired for 230 v.a.c. operation.

3. The PSU-6 is conservatively rated at 20 amps at 13.6 volts for intermittant operation and 10 amps semi-continuous operation.

   * Intermittant operation: Operated under a cycle of one minute on and four minutes off.

   * Semi-continuous: Eight hours continuous in 24 hours.

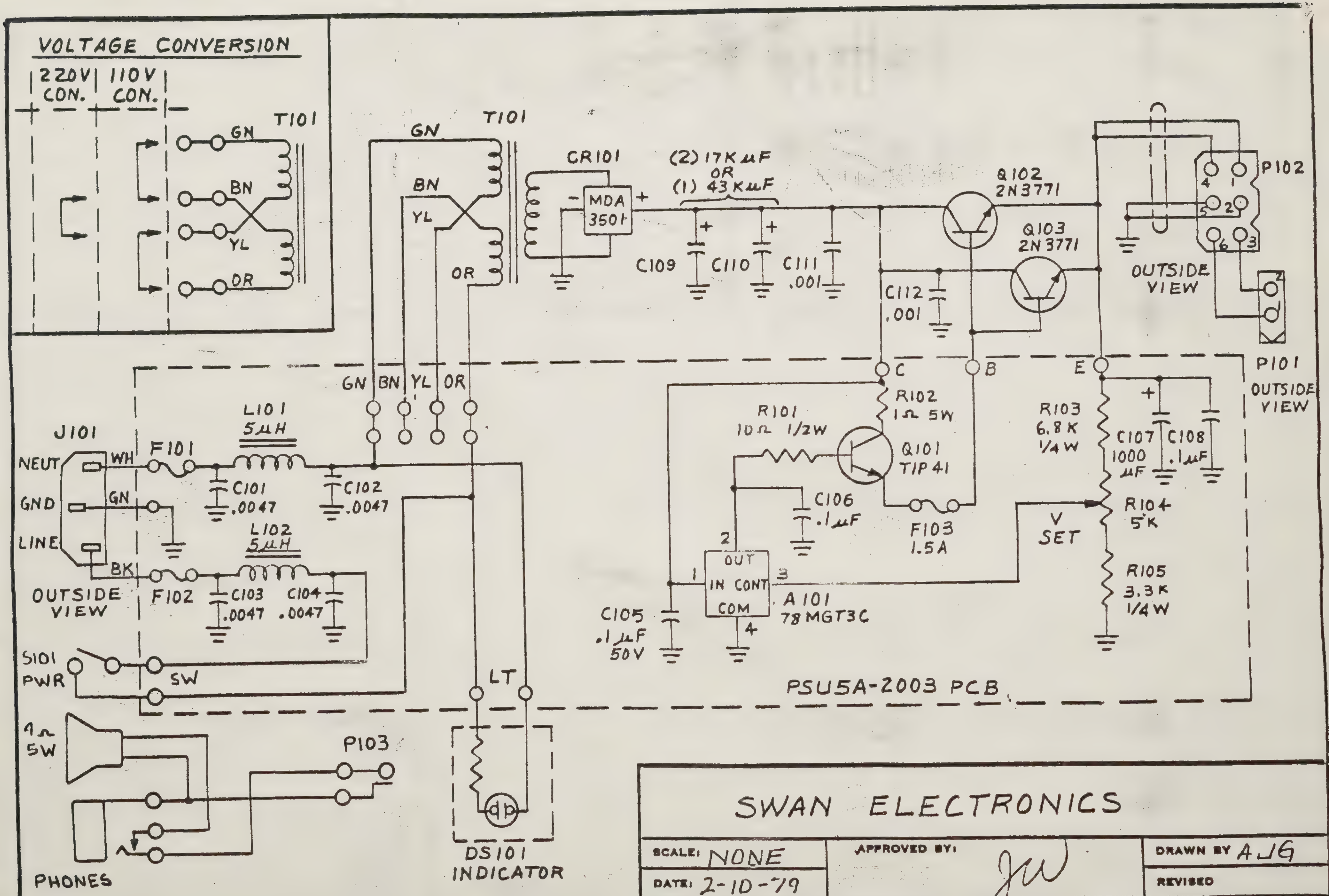
   The output voltage is factory set to 13.6 v.d.c. and is adjustable over a range of approximately 10 to 23 volts. The output regulation at full load is 3%, ripple is less than 1%. The heatsink is adequate for operation under the above operating conditions, however, under prolonged operation, the heatsink will reach temperatures too hot to touch. The supply should not be positioned so that circulation of air around the heatsink is inhibited. If operation is intended over a long period of time or under environmental conditions of high ambient temperature, it is recommended that a small fan be used to direct air against the heatsink.

4. The PSU-6 is styled to match the Swan transceivers and is complete with A.C. line cord, D.C. power cable with connector for the transceiver. The front panel contains ON-OFF switch and indicator light. The PSU-6 also has a speaker for external speaker operation, and a jack for low impedance headphones.

5. A two conductor connector at the end of the D.C. power cable is provided to power certain accessories (i.e., Pilot Lights of Antenna Tuner) and is fused by fuse in transceiver.

Specifications:

Average Power Consumption:	117 watts receive 700 watts transmit
Input:	117/230 VAC 50-60 Hz
Dimensions:	9 3/4" W x 5 3/4" H x 10 3/4 D
Weight:	19 lbs, 12 oz.; 8.96 Kg



SWAN ELECTRONICS

SCALE: NONE

DATE: 2-10-79

APPROVED BY: JW

DRAWN BY AJG

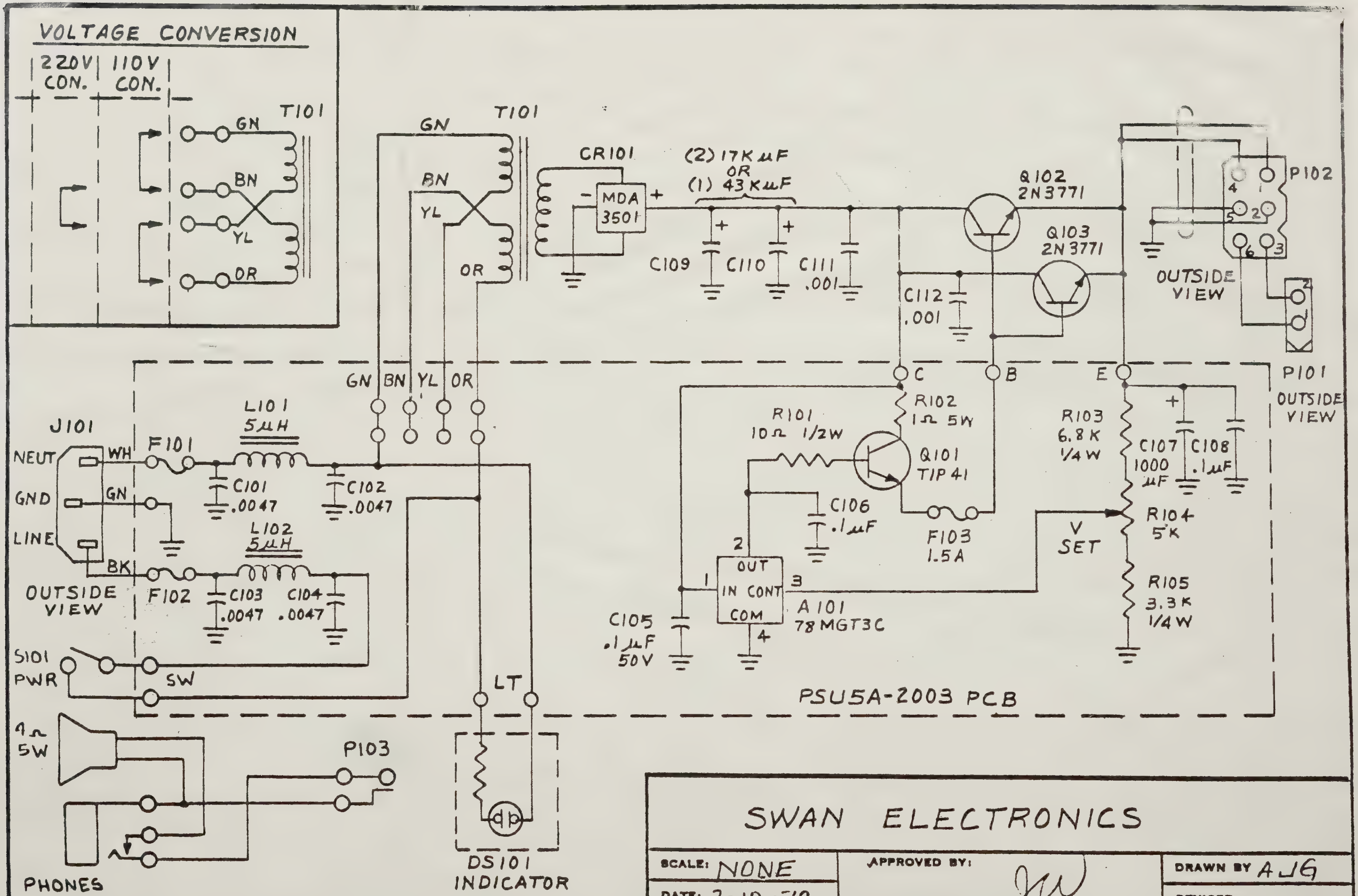
REVISED

SCHEMATIC-POWER SUPPLY (PSU-6)

DRAWING NUMBER
PSU6A-2401

'1. FUSES: F101, F102: 110V = 6A
220V = 4A

NOTES:



SWAN ELECTRONICS

SCALE: NONE

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REVISED

SCHEMATIC-POWER SUPPLY (PSU-6)

DRAWING NUMBER PSU6A-2401

1. FUSES: F101, F102: 110V = 6A
220V = 4A

NOTES:

SERVICE
MANUAL
ADDENDUM

ASTRO-103
TRANSCEIVER

MODIFICATION
TO THE
ASTRO-102BX
TRANSCEIVER

CUBIC COMMUNICATIONS,INC.
A member of the Cubic Corporation family of companies

Contents

Section One
General Information....................1-1

Section Two
Specifications.........................2-1

Section Three
Receiver Theory of Operation..........3-1

Section Four
Transmitter Theory of Operation.......4-1

Section Five
Oscillator Theory of Operation........5-1

Section Six
Ancillary Circuits Theory of Operation..............................6-1

Section Seven
Servicing Hints........................7-1

Section Eight
Servicing Data.........................8-1

Section Nine
Parts Lists............................9-1

Since Cubic Communications, Inc. is constantly striving to improve its products, it reserves the right to make changes without notice or obligation.

The information contained in this document is the property of Cubic Communications, Inc. in conjunction with operating and maintaining the ASTRO-103 Transceiver. This information may not be copied, reprinted, or disclosed to a third party, either wholly or in part, without the written consent of Cubic Communications,Inc.

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CUBIC
COMMUNICATIONS,INC.
A member of the Cubic Corporation family of companies.

To the Service Technician:

The ASTRO-103 is an updated version of the field proven ASTRO-102BX, adding the new amateur bands designated during the 1979 World Administrative Radio Conference (WARC). The new bands are added by substitution of the bandswitch and lowpass filter assembly and by the addition of preselector and local oscillator sections corresponding to the new bands at 30, 17 and 12 meters.

In addition to the new bands, an RTTY mode has been added to the transceiver that enables the use of the optional 6 pole narrow bandwidth crystal filter while operating on the lower sideband, such as is normal in amateur radio teletype using audio frequency shift keying. Some minor changes were made in the PTO mode switching sequence in response to customer suggestions.

All other features of the ASTRO-102BX are incorporated into the ASTRO-103. It is possible to convert an ASTRO-102BX transceiver into an ASTRO-103 by the addition of the new boards and some minor modifications to the existing boards following the step by step procedure written for the ASTRO-103 conversion kit. You can perform this conversion for your customers by ordering the conversion kit.

This manual is written as an addendum to the ASTRO-102BX Service Manual to emphasize the similarity between the two transceivers. This manual is organized in exactly the same sequence as the ASTRO-102BX Service Manual, with only the significant differences between the two transceivers indicated in the addendum.

You will note that the Swan name has been dropped and only the Cubic Communications name has been retained. The organization remains dedicated to the goal of serving the needs of our customers and our service technicians throughout the world.

Sincerely,

CUBIC COMMUNICATIONS, INC.

Jess C. Wright

Jess C. Wright
President

JCW/lb

305 Airport Rd. • Oceanside, Ca. 92054 • Telephone (714) 757-7525 Telex: 695435; ANS BK: CUBICOM, OCEN.

General

Information

The Cubic Communications, Inc. (CCI) ASTRO-103 is similar to the Swan ASTRO-102BX and the CCI ASTRO-102BXA with the addition of the new bands at 30, 17, and 12 meters.

The old standard 2 pole narrow band crystal CW filter has been deleted and replaced with an optional 6 pole unit having a 6dB bandwidth of 400 Hz. Older model transceivers can be upgraded to use the new 6 pole narrow CW filter with the addition of an accessory circuit board. Newer model transceivers have space on the exciter circuit board to accommodate the optional filter.

The RTTY mode in the ASTRO-103 transceiver is useable only if the optional CW narrow filter is installed. This mode is basically LSB narrow with the position of the 400 Hz bandwidth filter controlled by the passband tuning control.

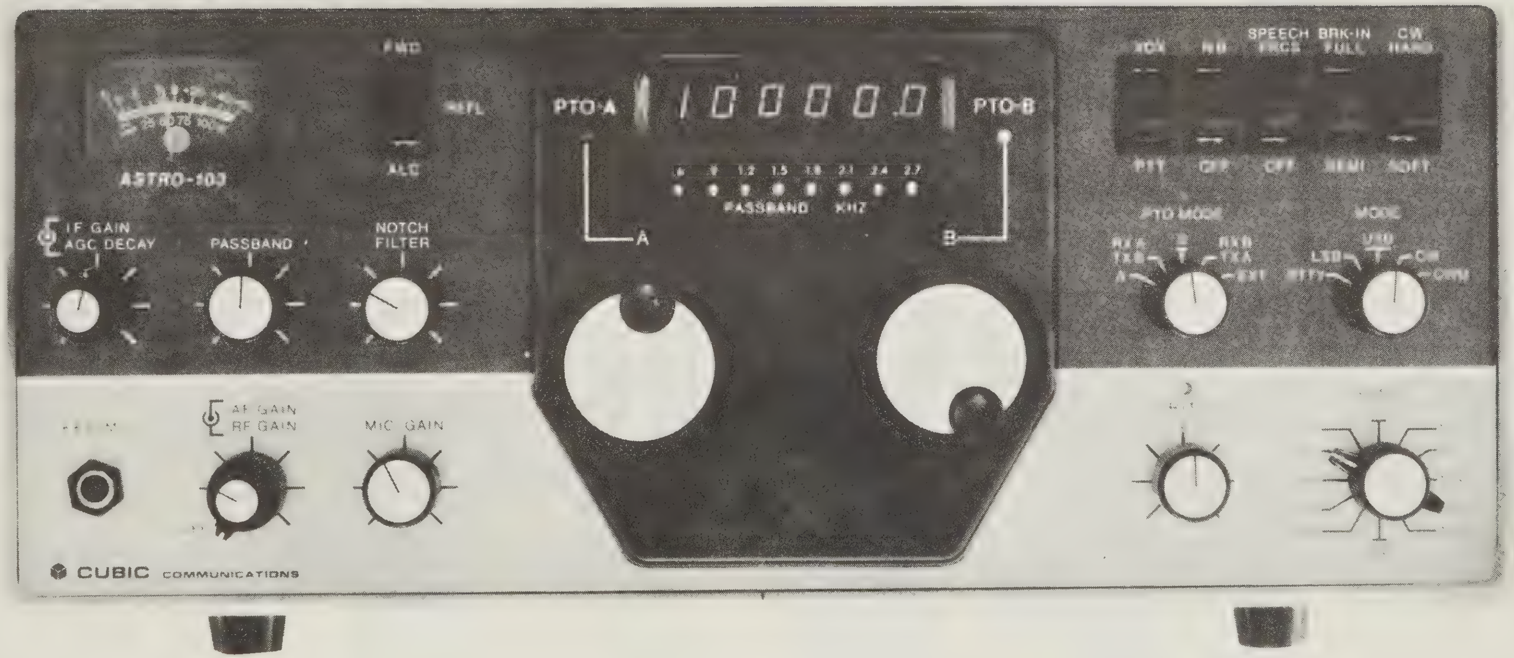

ASTRO-103
FWD
REFL
ALC
PTO-A
PTO-B
A
B
PASSBAND KHZ
.6
.9
1.2
1.5
1.8
2.1
2.4
2.7
VOX
NB
SPEECH PROC
BRK-IN FULL
CW HARD
PTT
OFF
OFF
SEMI
SOFT
PTO MODE
MODE
RX A TX B
A
B
RX B TX A
EXT
LSB
USB
CW
RTTY
IF GAIN
AGC DECAY
PASSBAND
NOTCH FILTER
AF GAIN
RF GAIN
MIC GAIN
CUBIC COMMUNICATIONS

FIGURE 5-1

Specifications

GENERAL

Frequency Coverage

For the ASTRO-103, add the following bands:

30M..............10.0-10.5 MHz
17M..............18.0-18.5 MHz
12M..............24.5-25.0 MHz

Modes of Operation:

For the ASTRO-103, add the following:

CW Narrow with the optional filter
RTTY (LSB Narrow) with the optional filter

Rear Panel:

For the ASTRO-103, add the following:

Separate Receiving Antenna:	RCA Jack
Receiver Antenna Selector:	Slide Switch

RECEIVER

Receiver Selectivity:

For the ASTRO-103, change CWN to read:

CWN and RTTY--400 Hz bandwidth 6 pole crystal filter in series with 8 pole SSB filter

Receiver Theory

Bandswitch and Low Pass Filters....................3-2
Preselector..3-2
CW Narrow Filter...................................3-3

BANDSWITCH AND LOWPASS FILTER BOARD

The ASTRO-103 is similar to the ASTRO-102BX in that one of six lowpass filters is switched into the circuit in both receive and transmit conditions. The filter sections are shared on some bands, however, with the 30 and 40 meter bands utilizing one section, the 17 and 20 meter bands utilizing a second section, and the 12 and 15 meter bands sharing a third section. The component values for each filter section are somewhat modified from the values used in the ASTRO-102BX.

Receiver antenna switching is performed by relay K101 when the common or normal receive antenna selection is made on the rear panel switch added in the ASTRO-103. A second relay has been added in the ASTRO-103 to allow for use of a separate receiving antenna. The contacts on this second relay connect the antenna side of C122 (in the ASTRO-102BX schematic) to the separate receiving antenna input jack on the rear panel. The antenna selection switch on the rear panel merely selects which relay is activated by transistor Q102. The drive to shunt transistor, Q101, has been slightly modified in the ASTRO-103 to eliminate any slight delay in the opening of either relay. Otherwise the operation of this circuitry is similar in both transceivers.

The band control wafer, S103, on the bandswitch has been modified to provide additional band control lines. In order to preserve the wiring in the ASTRO-102BX transceivers being upgraded to the ASTRO-103 configuration, the control information for the three additional bands has been brought out through a separate connector for use elsewhere in the transceiver.

Refer to the schematic diagram for the lowpass filter board provided with the manual addendum for additional information.

PRESELECTOR

The preselector section in the ASTRO-103 transceiver operates in a similar fashion as the preselector section in the ASTRO-102BX. A second preselector board (Auxiliary Preselector) is mounted above the original preselector board to accommodate the three new bands with a coaxial cable interconnection between the boards. Switching of selected bands is performed by the control lines originating at S103 as before. Refer to the schematic diagrams for the Preselector and Auxiliary Preselector boards provided with this addendum.

CWN FILTER

The 2 pole CWN filter provided with the ASTRO-102BX has been deleted in the ASTRO-103. A 6 pole filter is available as an option. On early model transceivers, the 6 pole filter is mounted on an auxiliary circuit board above the area formerly containing the 2 pole filter. Late model transceivers utilize a circuit board having provisions for drop-in mounting of the 6 pole filter.

Transmitter Theory

All transmitter circuitry in the ASTRO-103 is identical to that in the ASTRO-102BX with the exception of the Preselector Bandpass Filters. As in the ASTRO-102BX, the identical filters are used in both receive and transmit.

Oscillator Theory

Band Voltage Controlled Oscillators........................5-2
Local Oscillator Mixer.......................................5-2

BAND VOLTAGE CONTROLLED OSCILLATORS

The band voltage controlled oscillators in the ASTRO-103 transceiver are essentially identical to those in the ASTRO-102BX. An Auxiliary Local Oscillator board is mounted above the Local Oscillator board to provide space for the oscillators used on the added bands with a coaxial cable interconnection between the two boards to couple the oscillator output.

The programming of counters U102 and U103 is accomplished in a similar manner in the two transceivers except the programmable read only memory (PROM) U105 is modified for the ASTRO-103. Previously unused states in the PROM are programmed to the desired conditions corresponding to the new bands. Late model ASTRO-102BXA transceivers are shipped with the PROM corresponding to either the ASTRO-102BX or ASTRO-103 configurations. This part can be recognized by the CCI part number marked on the top of U105 -- if the number is 478-012 it is suitable for an ASTRO-103, if the number is 478-008 it is suitable only for an ASTRO-102BX.

Some changes are made in the steering of the band control lines as shown in the schematic diagram on page 5-2 of the ASTRO-102BX Service Manual. The diode matrix shown in this diagram is replaced with a similar matrix on the Auxiliary L.O. board with provisions for the three additional bands. The output of the matrix is then wired to U105 using a 5 wire cable.

The outputs of the D.C. Error Amplifiers are brought up to the Auxiliary L.O. board to control the oscillators. The 12 meter oscillator utilizes the 10 meter error line while the 30 and 17 meter oscillators utilize the 160-15 meter error line.

The schematic diagrams for the Auxiliary L.O. and the L.O. board with the ASTRO-103 modification are included with this manual. Refer to these schematic diagrams for additional details.

LOCAL OSCILLATOR MIXER

The Auxiliary L.O. board also contains a set of three bandpass filters to accommodate the L.O. signal for the three new bands. The circuits are similar to those on the L.O. board with circuit selection performed by the band control lines.

In addition to the band VCOs, diode matrix and the bandpass filters, the Auxiliary L.O. board contains circuitry and connectors to properly interconnect all modules utilizing the band control signals. Refer to the schematic diagrams and main interconnect diagram for complete details.

Ancillary Circuits

FREQUENCY COUNTER

The frequency counter circuitry in the ASTRO-103 is modified to accommodate primarily the 30 meter band. As can be seen from the description in the ASTRO-102BX Service Manual, the most significant digit in the frequency display is controlled by external gating operating on the band control lines. The control wiring to the frequency counter is intercepted on the Auxiliary L.O. board in order to add the additional gating for the 30, 17, and 12 meter bands.

The problem with the 30 meter band, however, is that the most significant digit would be a 1 above 10 MHz and a 0 below 10 MHz. An external gating signal could not perform this function. An Auxiliary Counter circuit board is thus installed internally to the counter enclosure to sense the above or below 10 MHz condition. The most significant digit is thus controlled correctly on the 30 and 20 meter bands. All other bands have the most significant digit controlled directly by the band control lines.

Servicing Hints

The ASTRO-103 represents an addition to the capability and features of the ASTRO-102BX but utilizes the same type of circuitry and components. Consequently, no change in the text of this section is required. The following additional note, however, can be applied to both the ASTRO-102BX and the ASTRO-103.

PASSBAND TUNING

An analysis of the filter response curves shown in the passband tuning section in the ASTRO-102BX Service Manual will show that one edge of the overall response curve will remain fixed in frequency while the opposite edge can be varied by using the PASSBAND TUNING control. A clockwise rotation of the control from the mid position will hold the upper edge fixed while varying the lower edge. Conversely, a counter-clockwise rotation of the control will hold the lower edge fixed while varying the upper edge. In this manner, the overall bandwidth can be varied and can be used to eliminate interference near one corner or the other of the filter response curve.

A comparison with amateur transceivers from other manufacturers will sometimes reveal that the function called passband tuning by some manufacturers would be better described as IF shift. The function is sometimes known by that name. The IF shift function merely moves the same filter bandwidth over a small range of frequencies. Although an interfering signal occuring near one edge of the normal response band could be eliminated with this method, the shift of the overall band could introduce interference from signals near the opposite edge. Note that the passband tuning function, as implemented in the ASTRO-102BX and ASTRO-103 transceivers, requires a second crystal filter, two mixers and a front panel controllable oscillator, while the much less effective IF shift function only requires a method of varying the beat frequency oscillator and tuning oscillator together from a front panel control.

Servicing Data

Display Frequency Accuracy Check..........................8-2

Alignment Procedure
    Band VCO..8-2
    Preselector and L.O. Bandpass Filters.................8-2

Layouts/Schematics
    Auxiliary Preselector Board.......................8-6, 8-7
    103 Low Pass Filter/Band Selector Board...........8-8, 8-9
    Auxiliary L.O. Board (Layout)...................8-10, 8-11
    Auxiliary Counter Board.........................8-12, 8-13
    103 Front Panel Right Board.....................8-14, 8-15
    New Exciter Board...............................8-16, 8-17

Main Chassis/Interconnect Schematic......................8-18

As the ASTRO-103 represents an addition to the ASTRO-102BX, most of the servicing data remains unchanged. Only the additions or modification to the data required by the ASTRO-103 will be provided in this addendum.

DISPLAY FREQUENCY ACCURACY

In paragraph 3, add ".....and RTTY modes." after "LSB"

CWN TEST

The same test used for the CWN function can be repeated in the RTTY mode by substituting LSB for CW and RTTY for CWN. The RF generator frequency must be readjusted for the desired tone frequencies and will be below the indicated frequency in LSB and RTTY modes while it will be above the indicated frequency in CW and CWN modes.

BAND VCO ADJUSTMENT

The adjustments of the band VCOs for the 30, 17, and 12 meter bands are similar to the other bands except that the adjustment is on the Auxiliary L.O. board. Of course, this board must be removed in order to gain access to the lower board adjustments (if required).

Adjust the coils on the Auxiliary L.O. board according to the following chart only if required and with the board mounted in position:

BAND	COIL	E.L.	VOLTAGE
30	T203	160-15	3.75 ±.25
17	T202	160-15	3.75 ±.25
12	T201	12,10	3.75 ±.25

PRESELECTOR AND L.O. BANDPASS FILTER ADJUSTMENT

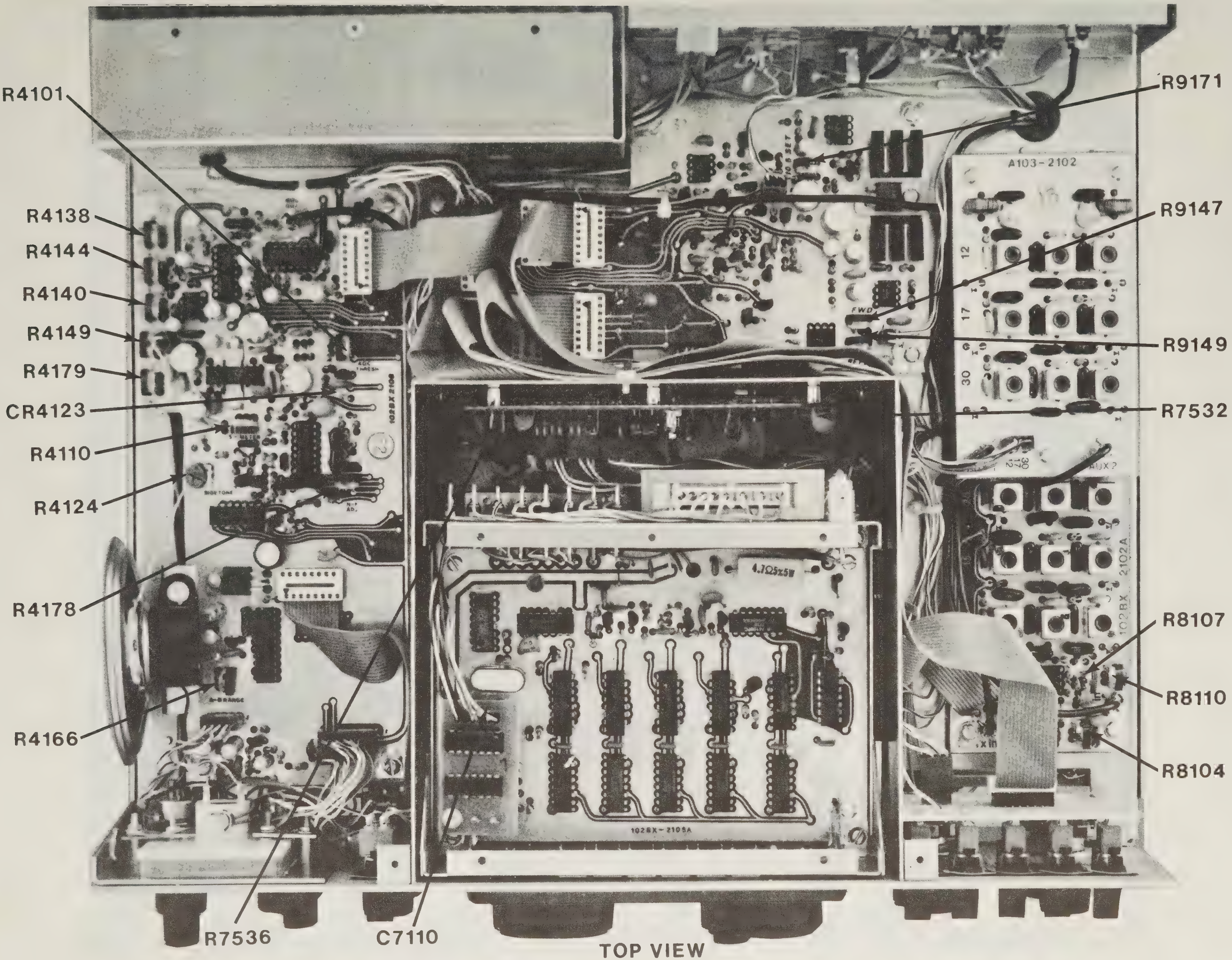
The bandpass filters on the Auxiliary Preselector and on the Auxiliary L.O. boards should be adjusted only if absolutely necessary and then only for those bands that require alignment. The auxiliary boards must be removed to allow access to the main board adjustments and then reinstalled to all for realignment of the auxiliary board transformers. The test points and method of test is identical for both the original ASTRO-102BX configuration and for the ASTRO-103 additions.

AUXILIARY PRESELECTOR BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHz	CENTER FREQ	COILS
30	10.0-10.5	10.250	T207, T208, T209
17	18.0-18.5	18.250	T204, T205, T206
12	24.5-25.0	24.750	T201, T203, T203

AUXILIARY L.O. BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHz	CENTER FREQ	COILS
30	19.0-19.5	19.250	T210, T211, T212
17	27.0-27.5	27.250	T207, T208, T209
12	33.5-34.0	33.750	T204, T205, T206



TOP VIEW
FIGURE 8-1

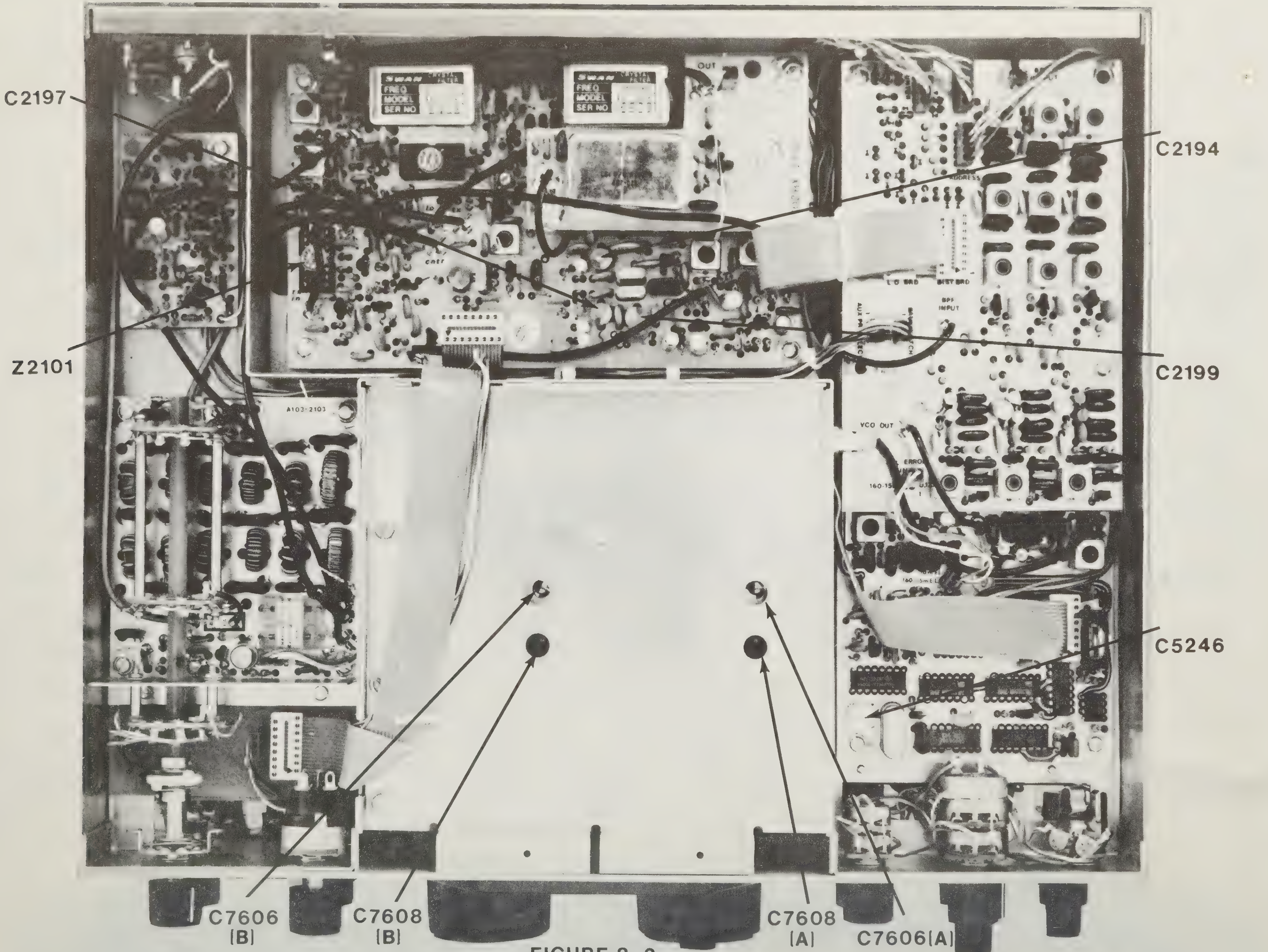


FIGURE 8-2
BOTTOM VIEW

Parts

Low Pass Filter/Band Switch Board (3000 Series)..............9-2
Auxiliary Local Oscillator Board (5200 Series)..............9-3
Front Panel Right Board (7800 Series)........................9-2
Auxiliary Preselector Board (8000 Series)...................9-4

Only parts unique to the ASTRO-103 are listed in this section. When refering to the parts lists for the ASTRO-102BX, the Lowpass Filter/ Band Switch and the Front Panel, Right boards are replaced in their entirety. The Auxiliary Preselector, Auxiliary L.O. and Auxiliary Counter boards are new boards used only in the ASTRO-103.

LOWPASS FILTER/BANDSWITCH BOARD (3000 Series)

Changes for the ASTRO-103 Only

Circuit Symbol	Cubic Part Number	Description
K3102	111-036	Relay, Reed SPST
L3107	027-081	Coil, RF, 1.15 uH, 14T (30, 40 Meters)
L3108	027-081	Coil, RF, 1.15 uH, 14T (30, 40 Meters)
L3111	027-098	Coil, RF, 6.4 uH, 33T (160 Meters)
L3112	027-098	Coil, RF, 6.4 uH, 33T (160 Meters)
S3101	171-205	Switch Wafer
S3102	171-205	Switch Wafer
S3103	171-205	Switch Wafer

FRONT PANEL, RIGHT BOARD (7800 Series)

Changes for the ASTRO-103 Only

Circuit Symbol	Cubic Part Number	Description
S7802	171-188	Switch, Mode

AUXILIARY L.O. BOARD (5200 Series)

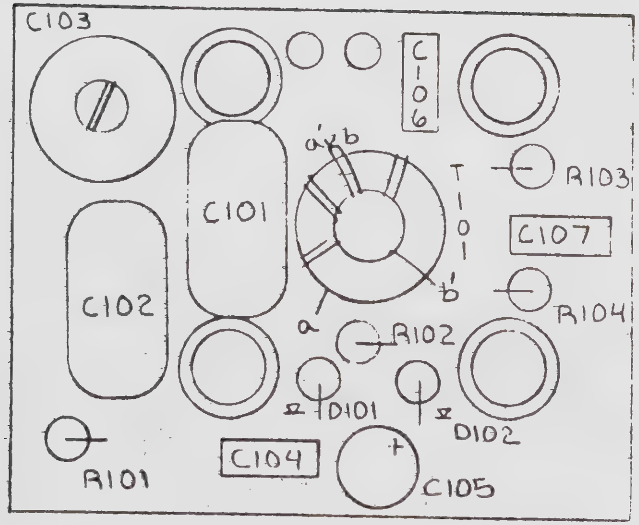
ASTRO-103 Only - Other parts similar to L.O. Board on ASTRO-102BX

Circuit Symbol	Cubic Part Number	Description
T5201	012-231	Transformer, RF 6T
T5202	014-066	Transformer, RF, 8T
T5203	012-255	Transformer, RF, 9T
T5204	014-066	Transformer, RF, 8T
T5205	014-067	Transformer, RF, 14T
T5206	014-066	Transformer, RF, 8T/12T
T5207	014-066	Transformer, RF, 8T
T5208	014-067	Transformer, RF, 14T
T5209	014-066	Transformer, RF, 8T/12T
T5210	014-066	Transformer, RF, 8T
T5211	014-067	Transformer, 14T
T5212	014-066	Transformer, RF, 8T/12T
L5201	027-059	RF Choke 1.5 uH
L5202	027-059	RF Choke 1.5 uH
L5203	027-060	RF Choke 1.8 uH
L5204	027-060	RF Choke 1.8 uH
L5205	027-062	RF Choke 2.7 uH
L5206	027-062	RF Choke 2.7 uH
U5105	476-012	PROM with code for ASTRO-103 Used on L.O. Board, replaces 476-008 (HM7603 or N82S123)

AUXILIARY PRESELECTOR BOARD (8200 Series)

ASTRO-103 Only - Other parts similar to Preselector Board on ASTRO-102BX

Circuit Symbol	Cubic Part Number	Description
L8201	012-269	Coil, RF 7 uH, 40T
L8202	012-269	Coil, RF 17 uH, 40T
C8218	089-028	Capacitor, Trimmer 4-30 pf
C8219	089-028	Capacitor, Trimmer, 4-30 pf
T8201	014-066	Transformer, RF, 8T/2T
T8202	014-067	Transformer, RF, 14T/3T
T8203	014-066	Transformer, RF 8T/2T
T8204	014-067	Transformer, RF 14T/3T
T8205	014-068	Transformer, RF 20T/3T
T8206	014-068	Transformer, RF 20T/3T
T8208	014-069	Transformer, RF, 32T/9T
T8209	014-068	Transformer, RF 20T/3T



.COMPONENT DESIGNATORS (7700 SERIES)
NOTES:

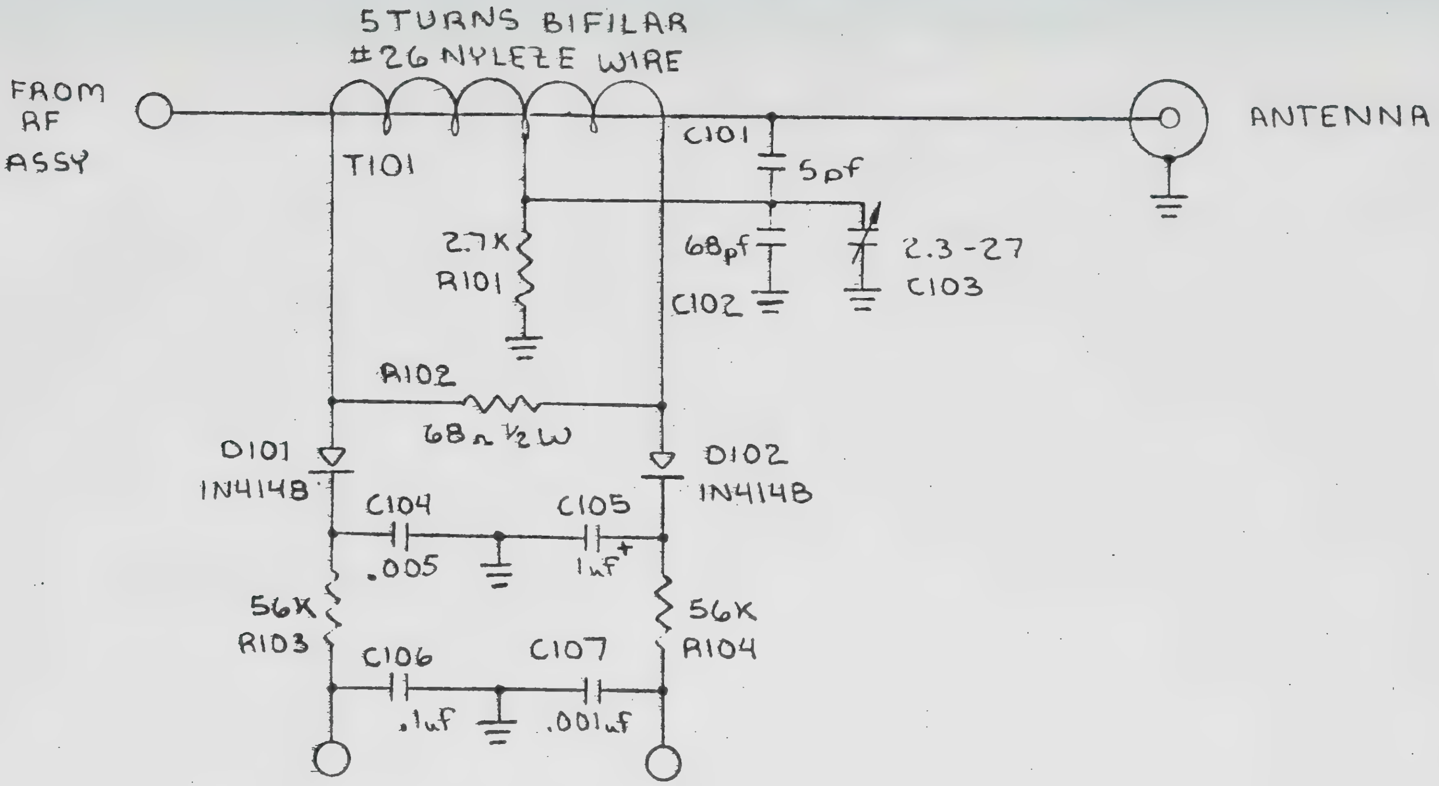
SWAN ELECTRONICS

SCALE: NONE
DATE: 1-11-79
APPROVED BY:
DRAWN BY C. DAVIDSON
REVISED

SWR / WATTMETER BRD LAYOUT

DRAWING NUMBER
0800-2303

LTR	DESCRIPTION	DATE	APP'D
A	CHANGED C106 FROM .001 TO .1uf CD 8-24-79	8-24-79	Jw



1. COMPONENT DESIGNATORS (7700 SERIES)
NOTES

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DATE: 1-11-79		REVISED
SWR/WATTMETER BRD SCHEMATIC		
		DRAWING NUMBER 0800-2403A

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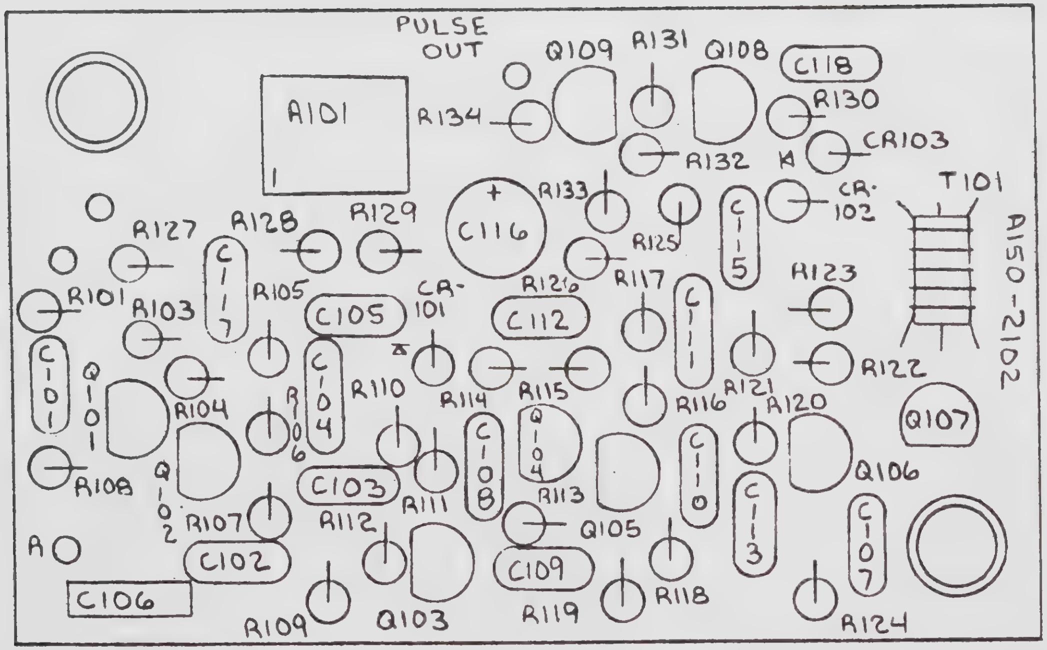

SWAN ELECTRONICS

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DATE: 1-12-79

APPROVED BY: JW

DRAWN BY C.DAVIDSON

REVISED

PROJ.

0176

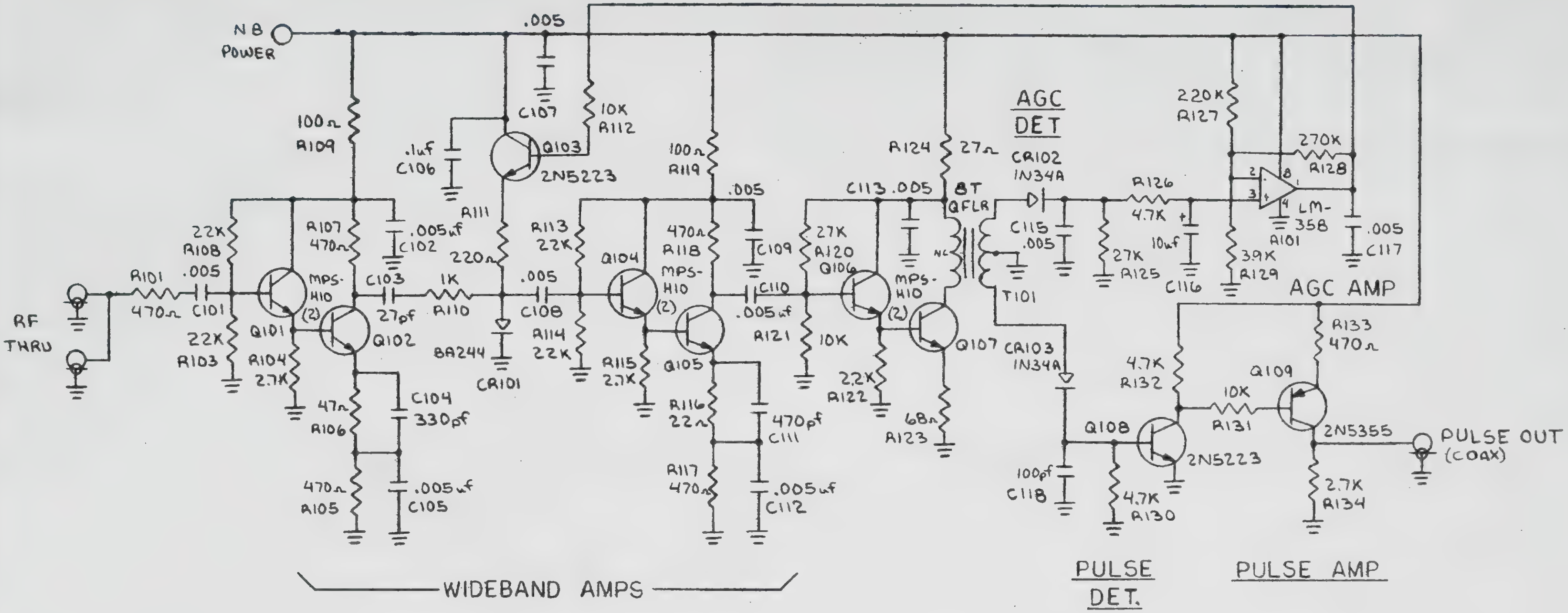
A150 NOISE BLANKER BRD LAYOUT

DRAWING NUMBER

0800-2304

1.COMPONENT DESIGNATORS (7400 SERIES)

NOTES



1. COMPONENT DESIGNATORS (7400 SERIES)
NOTES

SWAN ELECTRONICS

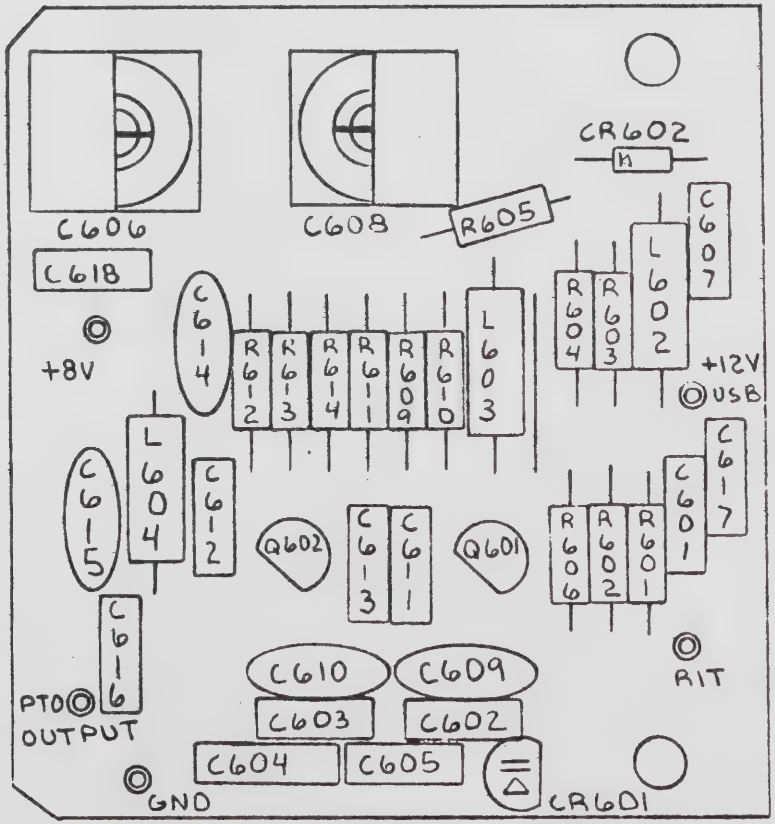
SCALE: NONE
DATE: 1-12-78
APPROVED BY: JW
DRAWN BY C. DAVID
REVISED

PROJ.
0176

A150 NOISE BLANKER BRD SCHEMATIC

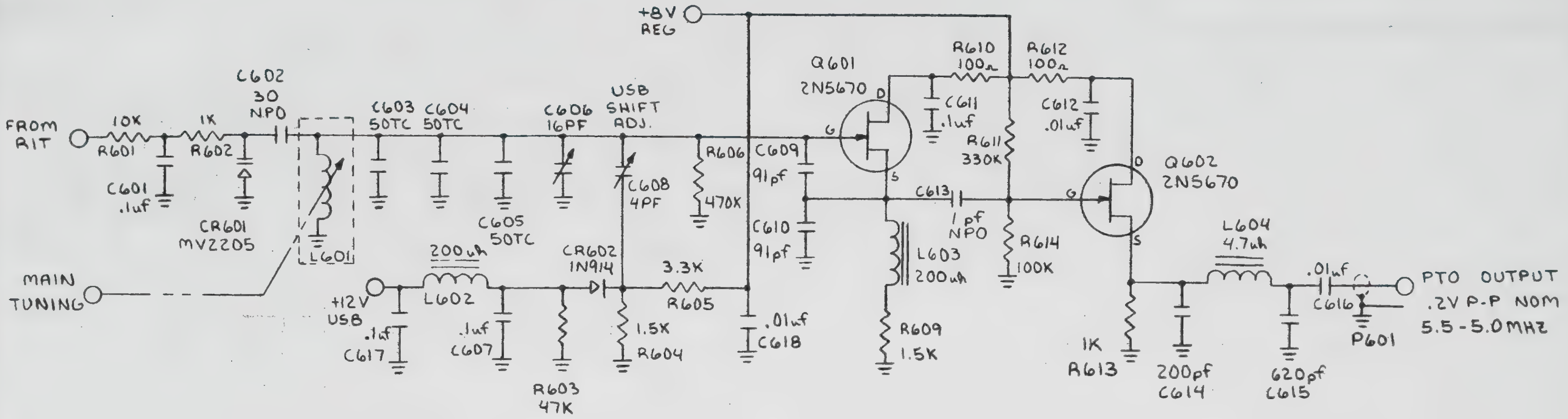
DRAWING NUMBER
0800-2404A

11X17 PRINTED ON NO. 1000H CLEARPRINT



SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY:	DRAWN BY C DAVIDSON
DATE: 9-14-78		REVISED
PTO BRD LAYOUT		
		DRAWING NUMBER 0800-2305

1. COMPONENT DESIGNATORS 7600 SERIES



1. COMPONENT DESIGNATORS 7600 SERIES

NOTES

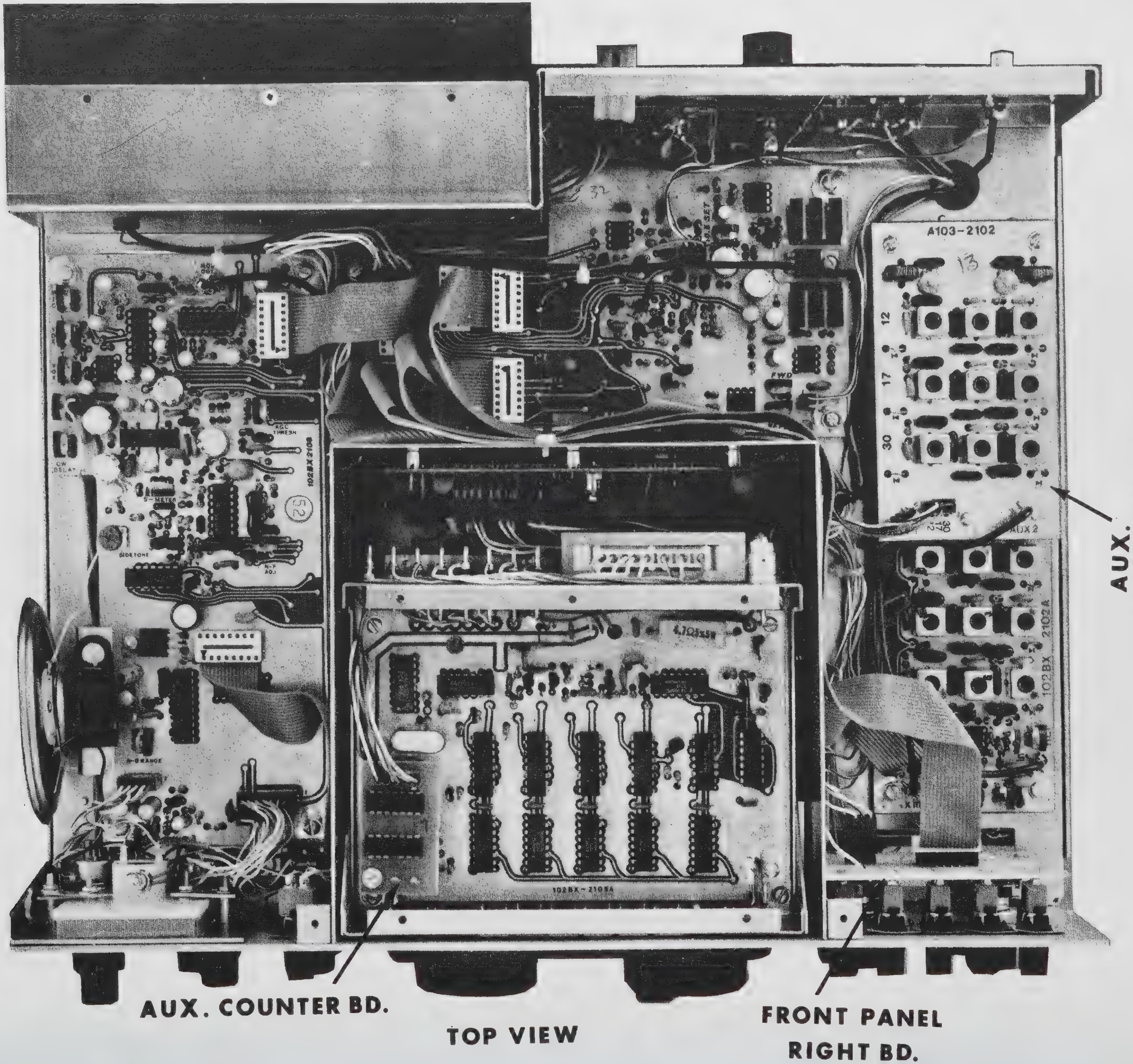
SWAN ELECTRONICS

SCALE: NONE
DATE: 9-22-78
APPROVED BY:
DRAWN BY C. DAVIDSON
REVISED

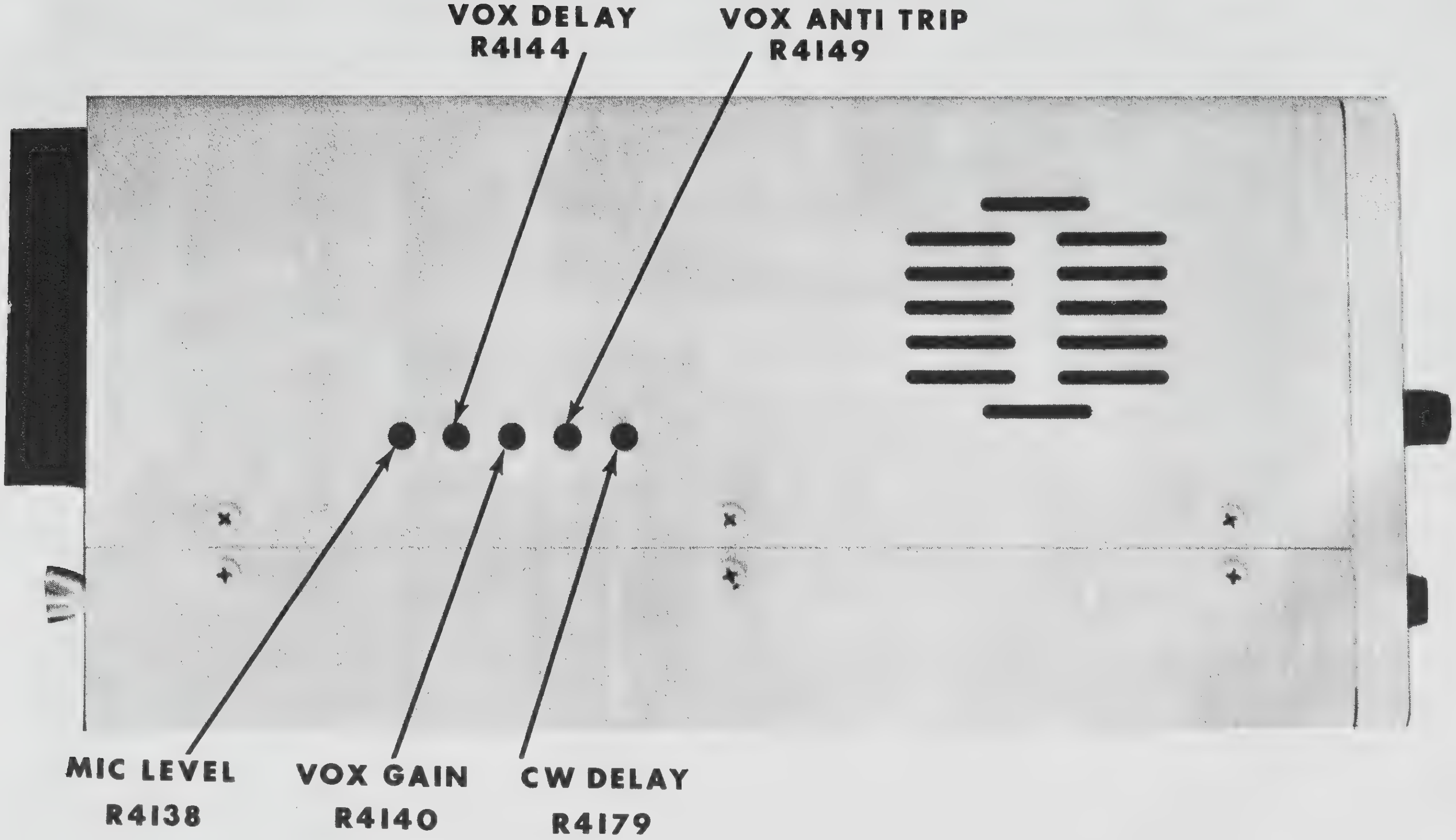
PTO BRD SCHEMATIC (038-188A)

DRAWING NUMBER 0800-2405

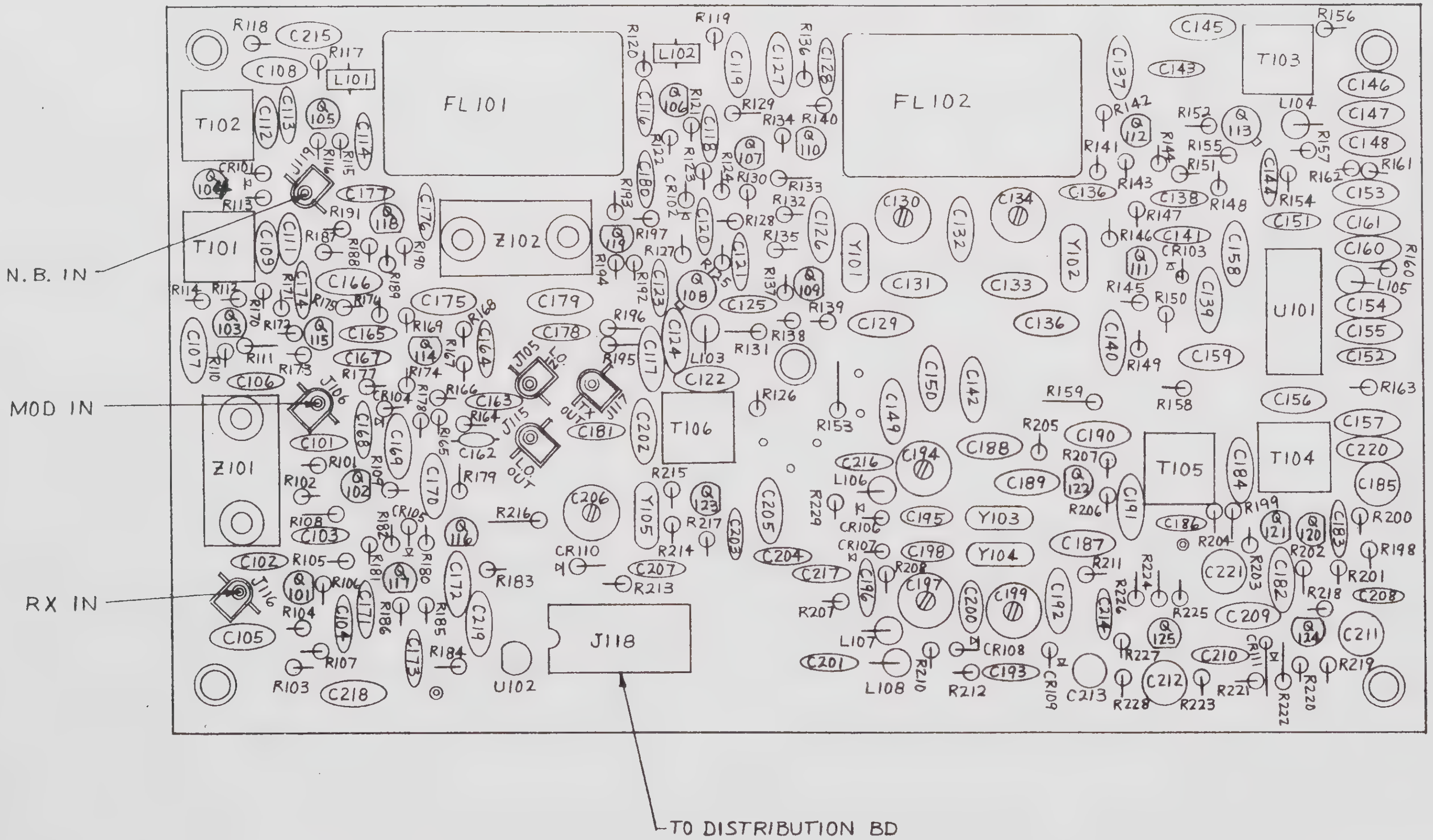
11X17 PRINTED ON NO. 1000H CLEARPRINT


AUX. PRESELECTOR BD.
AUX. COUNTER BD.
FRONT PANEL RIGHT BD.
A103-2102
102BX-2102A
102BX-2105A
102BX-2104

TOP VIEW



LEFT SIDE VIEW



2. COMPONENT DESIGNATORS 2000 SERIES.
1. REF. ELECT. ASSY 102BX-2007
REF. CIRCUIT BOARD 102BX-2107
REF. DRILL DWG 102BX-2204

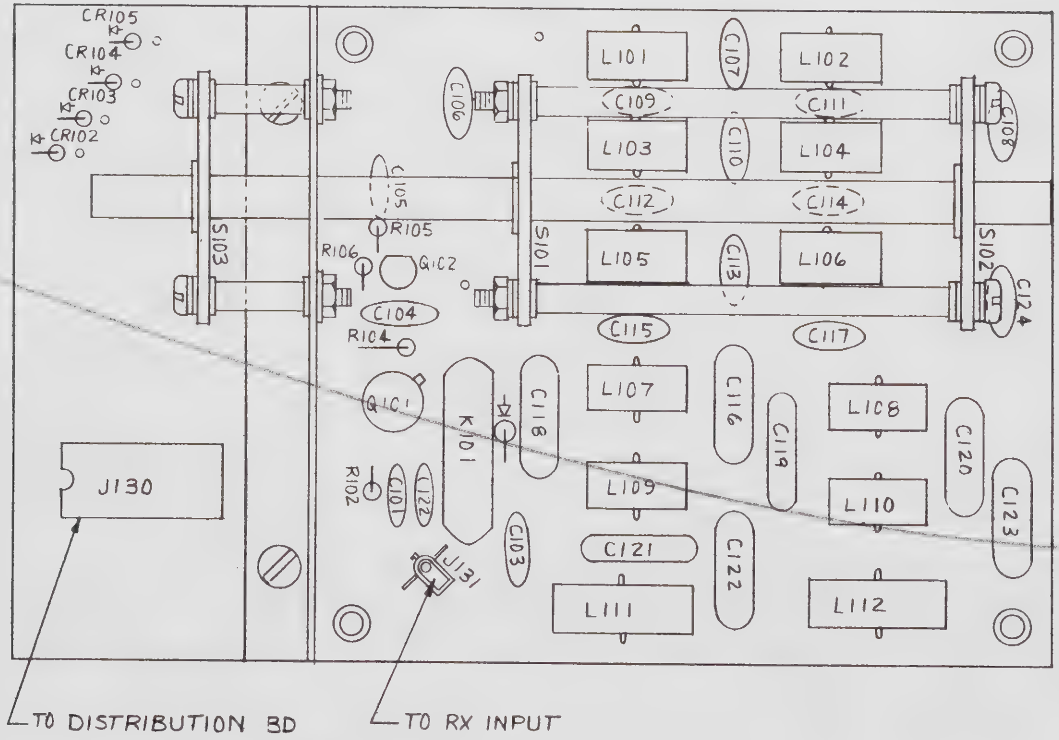
NOTES:

SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 9-24-79		REVISED

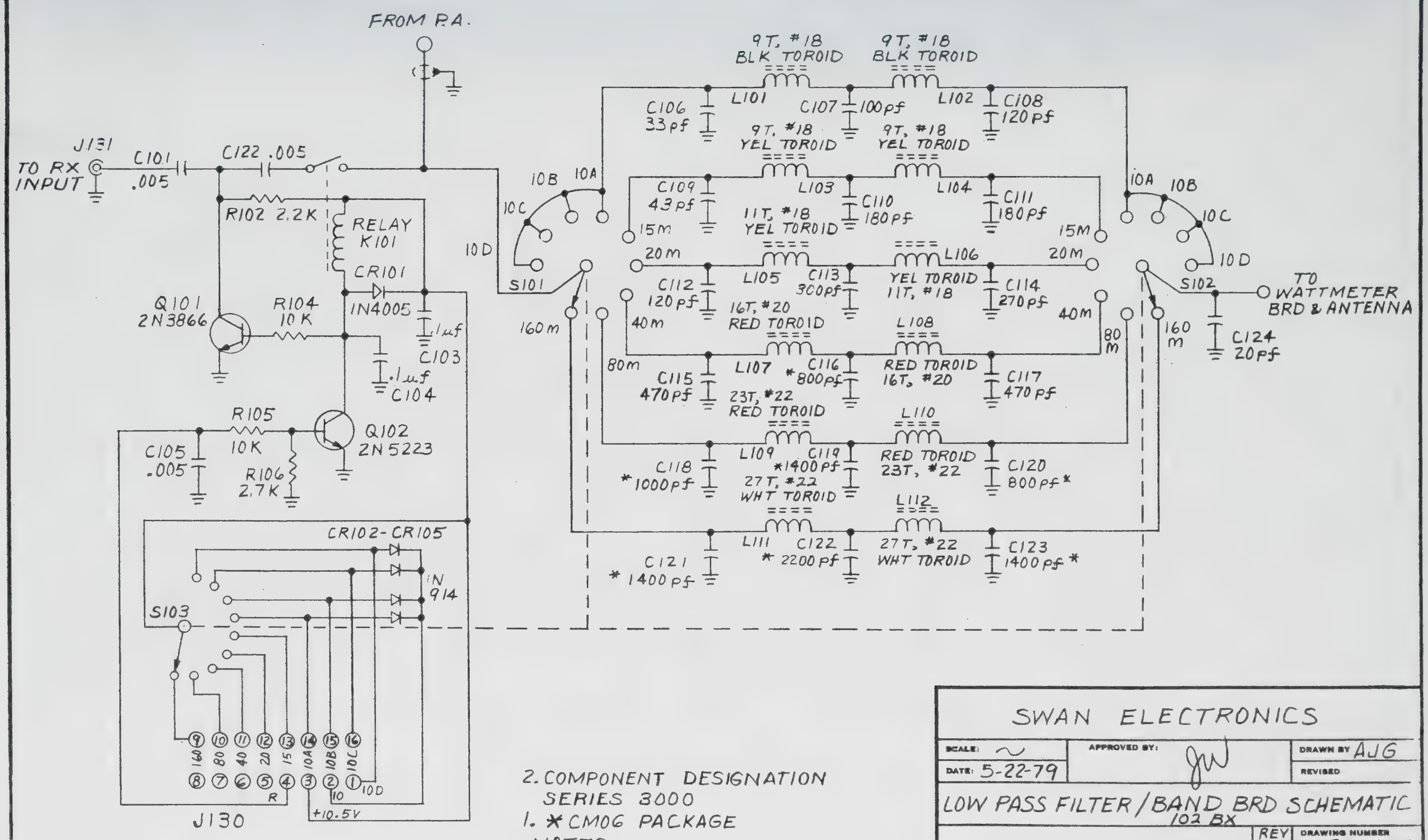
EXCITER BOARD LAYOUT

DRAWING NUMBER 102BX-2307



2. COMPONENT DESIGNATORS 3000 SERIES.
1. REF ELEC. ASSY 102BX-2003
   REF. CIRCUIT BD 102BX-2103
   REF. DRILL DWG 102BX-2203
NOTES:

SWAN ELECTRONICS
SCALE: ~
DATE: 9-18-79
APPROVED BY: JW
DRAWN BY AJG
REVISED
B'SWITCH & L.P. FILT. BD LAYOUT
DRAWING NUMBER 102BX-2303



2. COMPONENT DESIGNATION
SERIES 3000
1. * CM06 PACKAGE
NOTES:

SWAN ELECTRONICS

SCALE: ~
APPROVED BY: JW
DRAWN BY AJG
DATE: 5-22-79
REVISED

LOW PASS FILTER / BAND BRD SCHEMATIC
102 BX

REV
DRAWING NUMBER 102BX-2403

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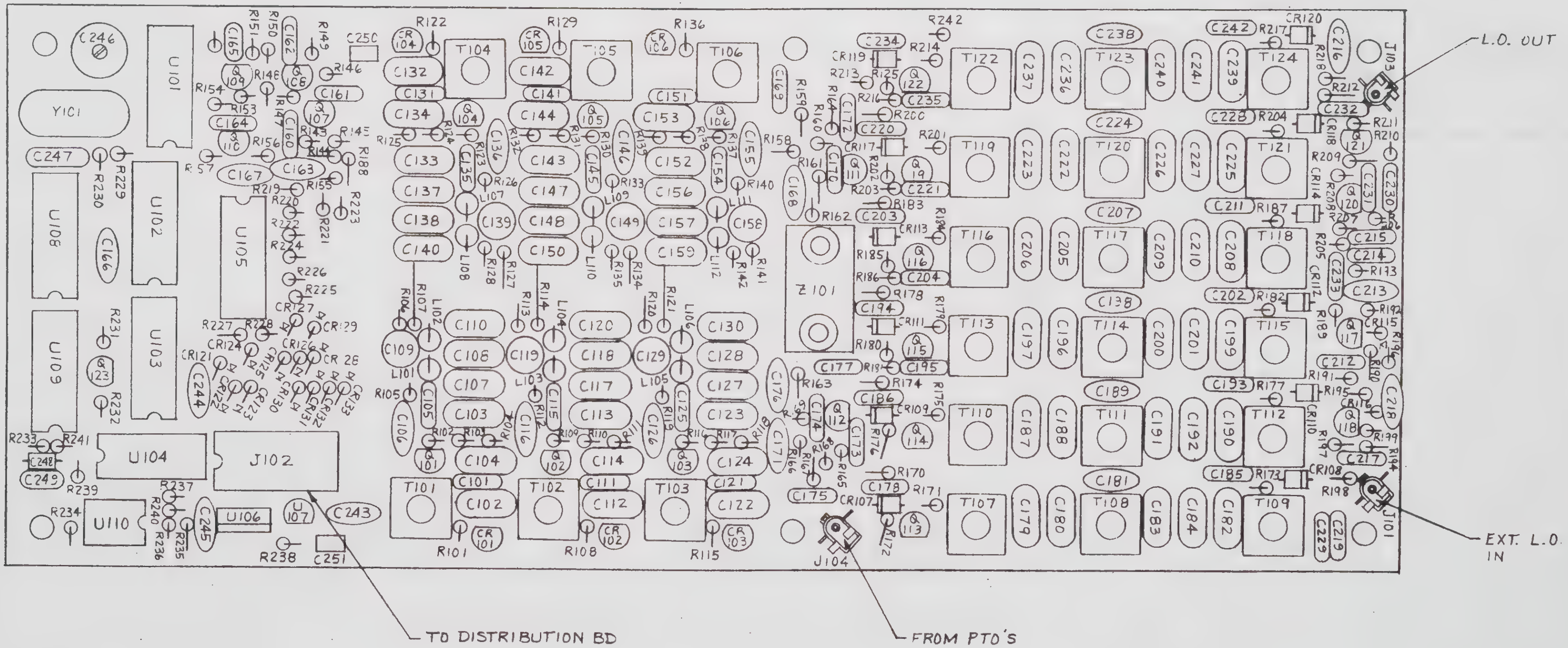


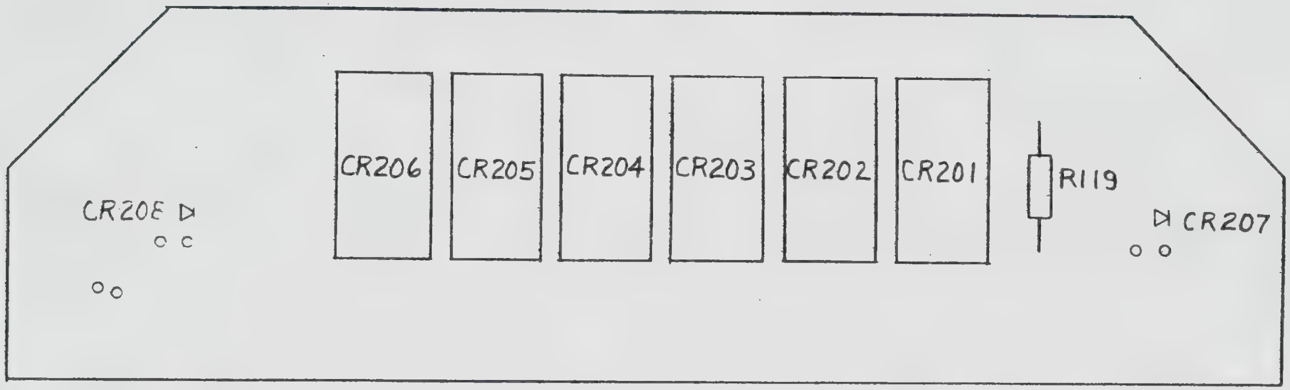
2. COMPONENT DESIGNATORS: 5000 SERIES

1. REF. ELECT. ASSY 102BX-2004
   REF. CIRCUIT BOARD 102BX-2104
   REF. DRILL DWG 102BX-2204

NOTES.

SWAN ELECTRONICS		
SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 7-30-79		REVISED
L.O. BOARD COMPONENT LAYOUT		
	REV.	DRAWING NUMBER 102BX-2304

18X24 PRINTED ON NO. 1000H-8 CLEARPRINT FADE-OUT



2. COMPONENT DESIGNATOR SERIES 7200
1. REF. ELECT. ASSY 102BX-2006
REF. CIRCUIT BD 102BX-2106
REF. DRILL DWG 102BX-2206
NOTES:

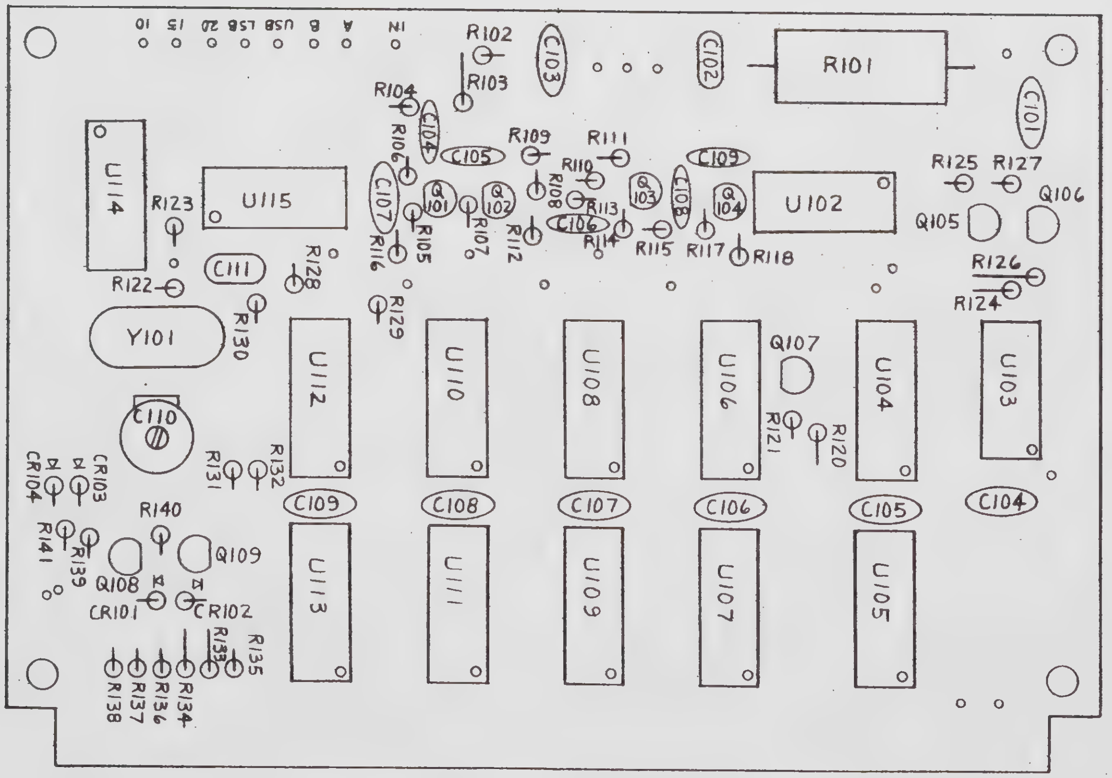
SWAN ELECTRONICS

SCALE: ∼	APPROVED BY: JW	DRAWN BY AJG
DATE: 9-19-79		REVISED

DISPLAY BOARD LAYOUT

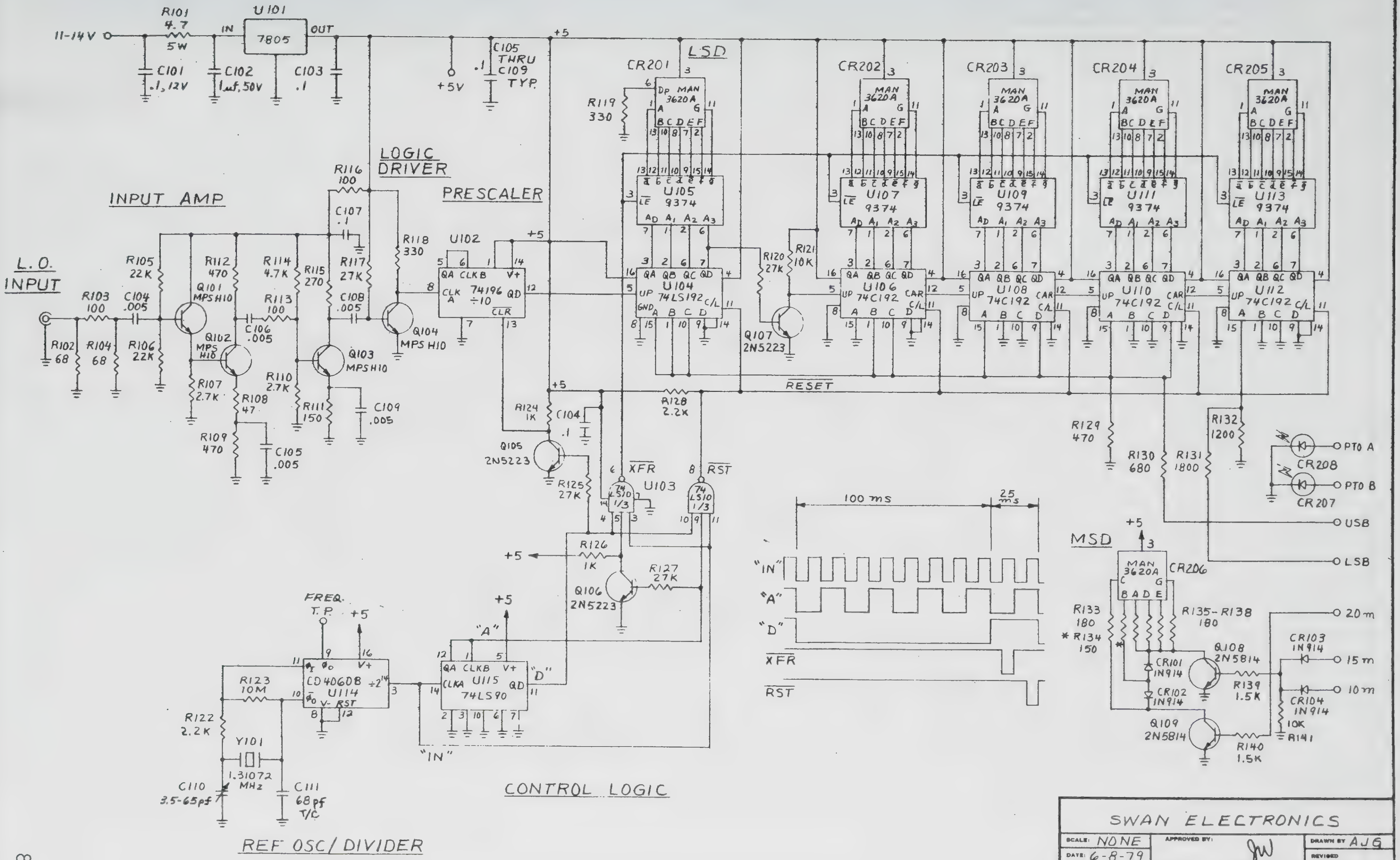
DRAWING NUMBER 102BX-2306

11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS 7100 SERIES
1. REF. ELEC. ASSY 102BX-2005
REF. CIRCUIT BD 102BX 2105
REF. DRILL DWG 102BX 2205
NOTES:

SWAN ELECTRONICS
SCALE: ~
APPROVED BY: JW
DRAWN BY AJG
DATE: 9-19-74
REVISED
COUNTER BOARD LAYOUT
DRAWING NUMBER 102BX-2305



2. DISPLAY REF. DESIGNATOR SERIES 7200
1. COUNTER REF. DESIGNATOR SERIES 7100
NOTES.

SWAN ELECTRONICS

SCALE: NONE	APPROVED BY: JW	DRAWN BY AJG
DATE: 6-8-79		REVISED
COUNTER / DISPLAY BD SCHEMATIC 102 BX		
	REV	DRAWING NUMBER 102BX-2405

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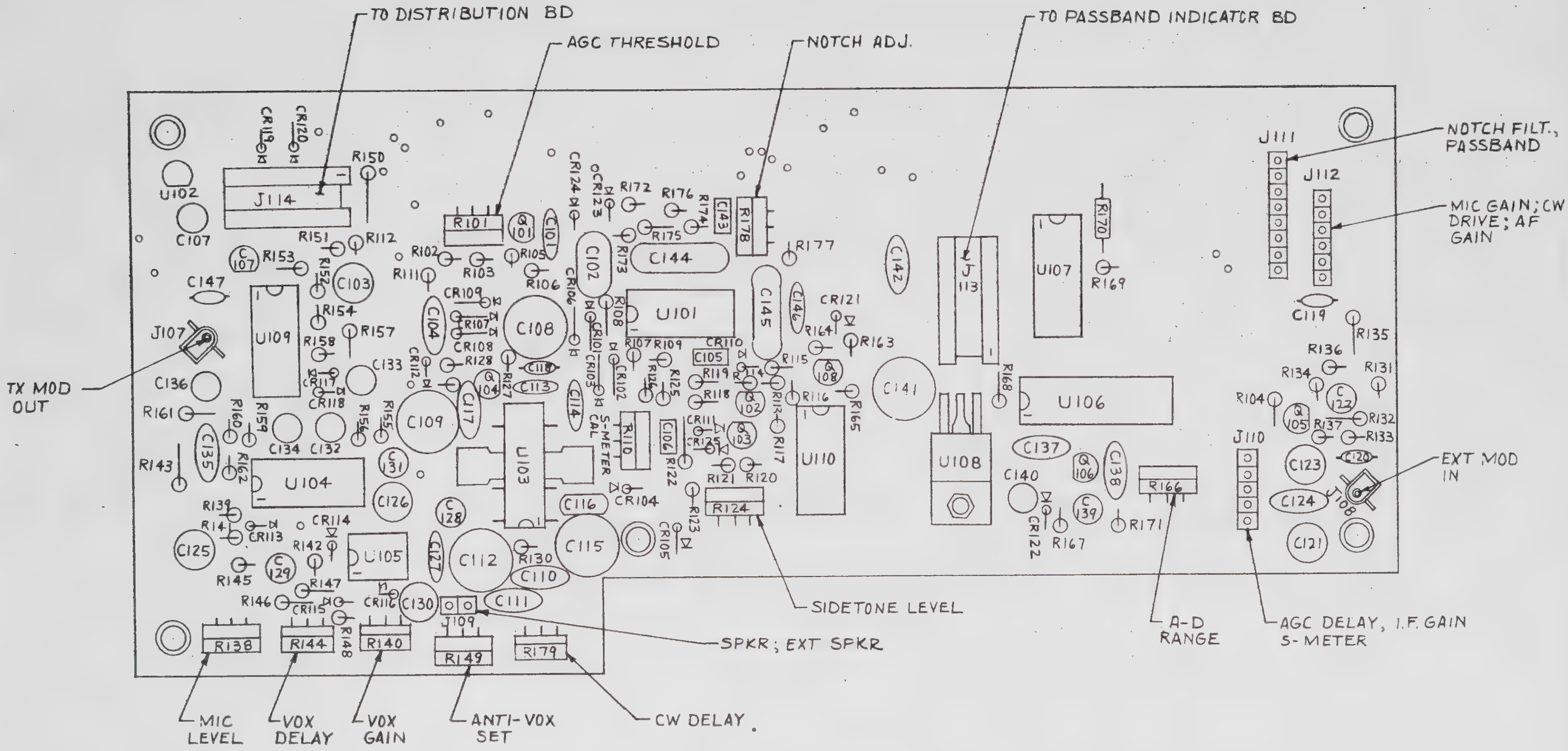
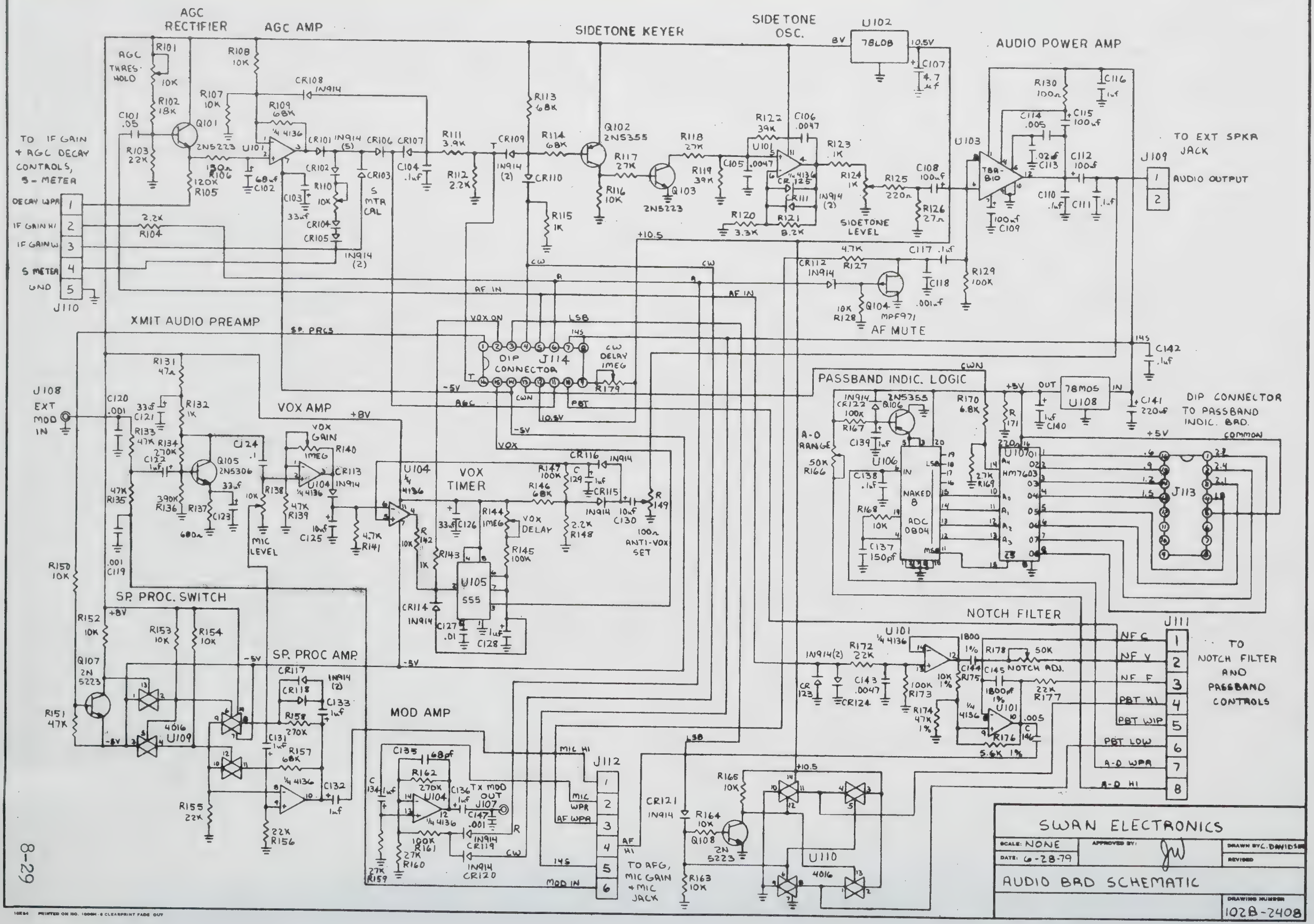


2. COMPONENT DESIGNATORS 4000 SERIES.
1. REF. ELECT. ASSY 102BX-2008
REF. CIRCUIT BOARD 102BX-2108
REF. DRILL DWG 102BX-2208
NOTES:

SWAN ELECTRONICS
SCALE: NONE
APPROVED BY: Jw
DRAWN BY AJG
DATE:
REVISED
AUDIO BOARD LAYOUT
DRAWING NUMBER 102BX-2308

NOTES: REF DESIGNATION SERIES 4000.



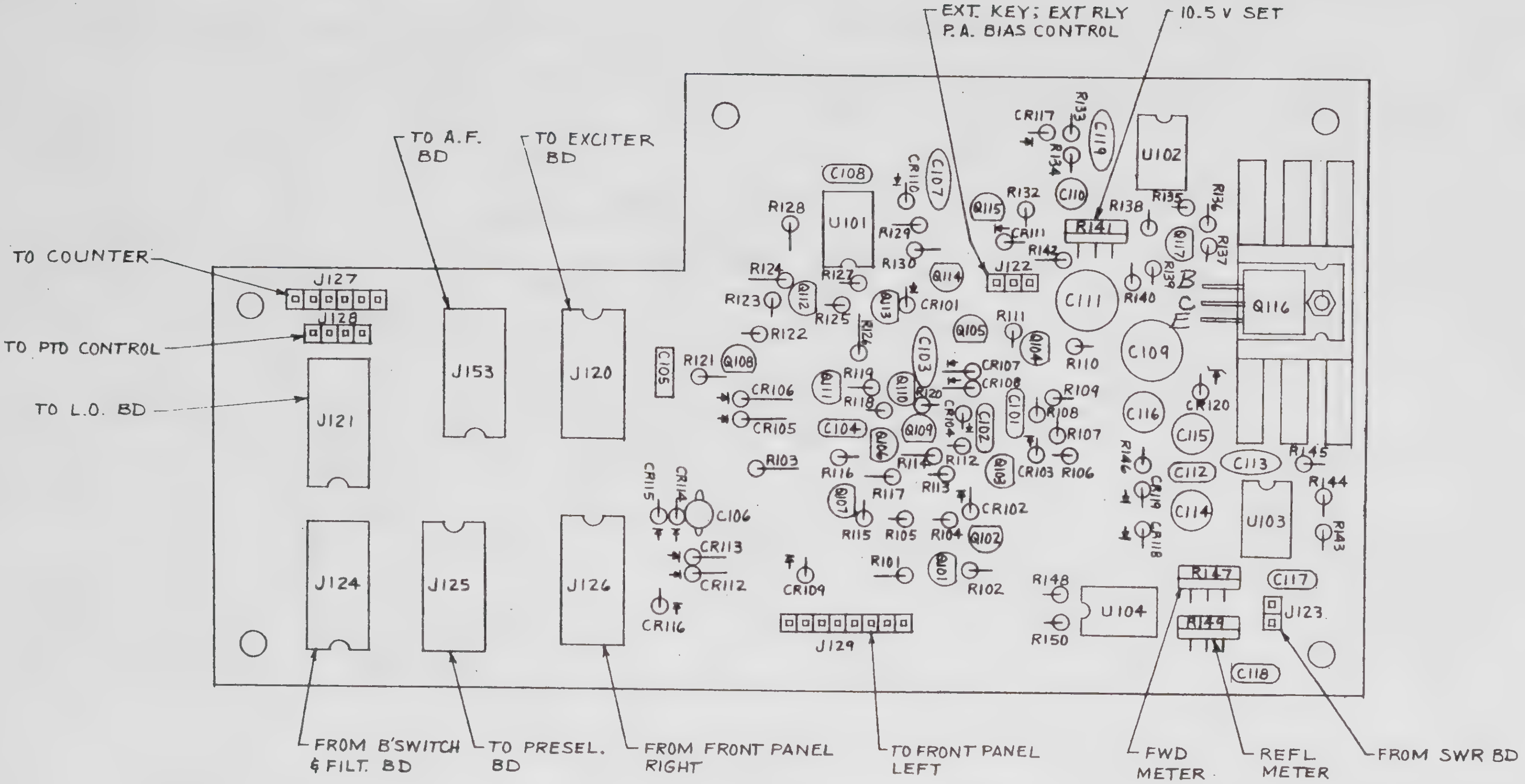
SWAN ELECTRONICS

SCALE: NONE
APPROVED BY: JW
DRAWN BY: C. DAVIDSON
DATE: 6-28-79
REVISED

AUDIO BRD SCHEMATIC

DRAWING NUMBER 102B-2408

PRINTED ON NO. 1000H-8 CLEARPRINT FADE OUT



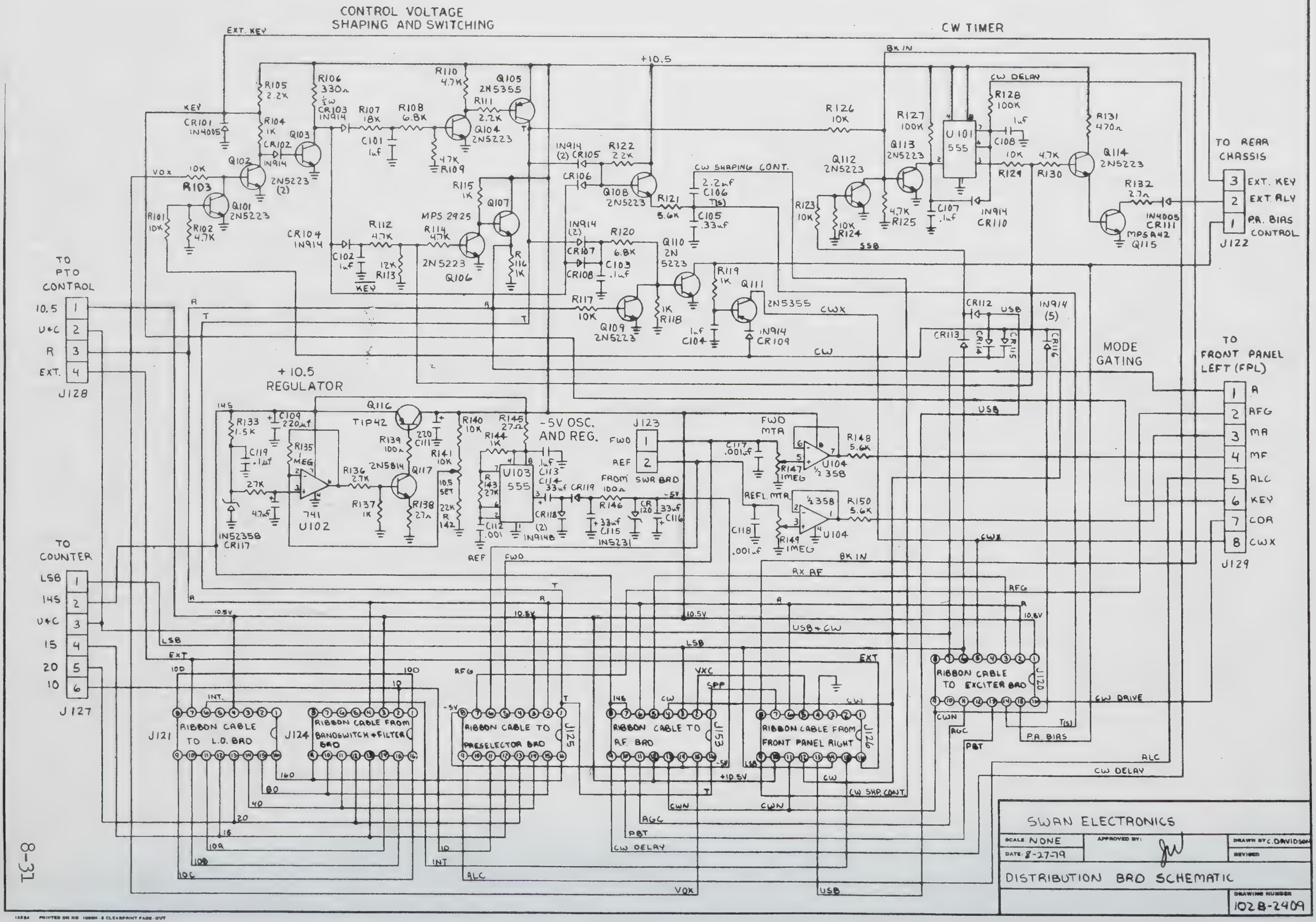
2. COMPONENT DESIGNATORS: 9000 SERIES

1. REF. ELECT. ASSY 102BX-2009
REF. CIRCUIT BD 102BX-2109
REF. DRILL DWG 102BX-2209
NOTES:

SWAN ELECTRONICS
SCALE: NONE
APPROVED BY: JW
DRAWN BY: AJG
REVISED
DATE: 7-25-79
DISTRIBUTION BOARD
COMPONENT LAYOUT
REV
DRAWING NUMBER 102BX-2309

PRINTED ON NO. 1000H-5 CLEARPRINT FADE OUT

COMPONENT DESIGNATORS:
9000 SERIES

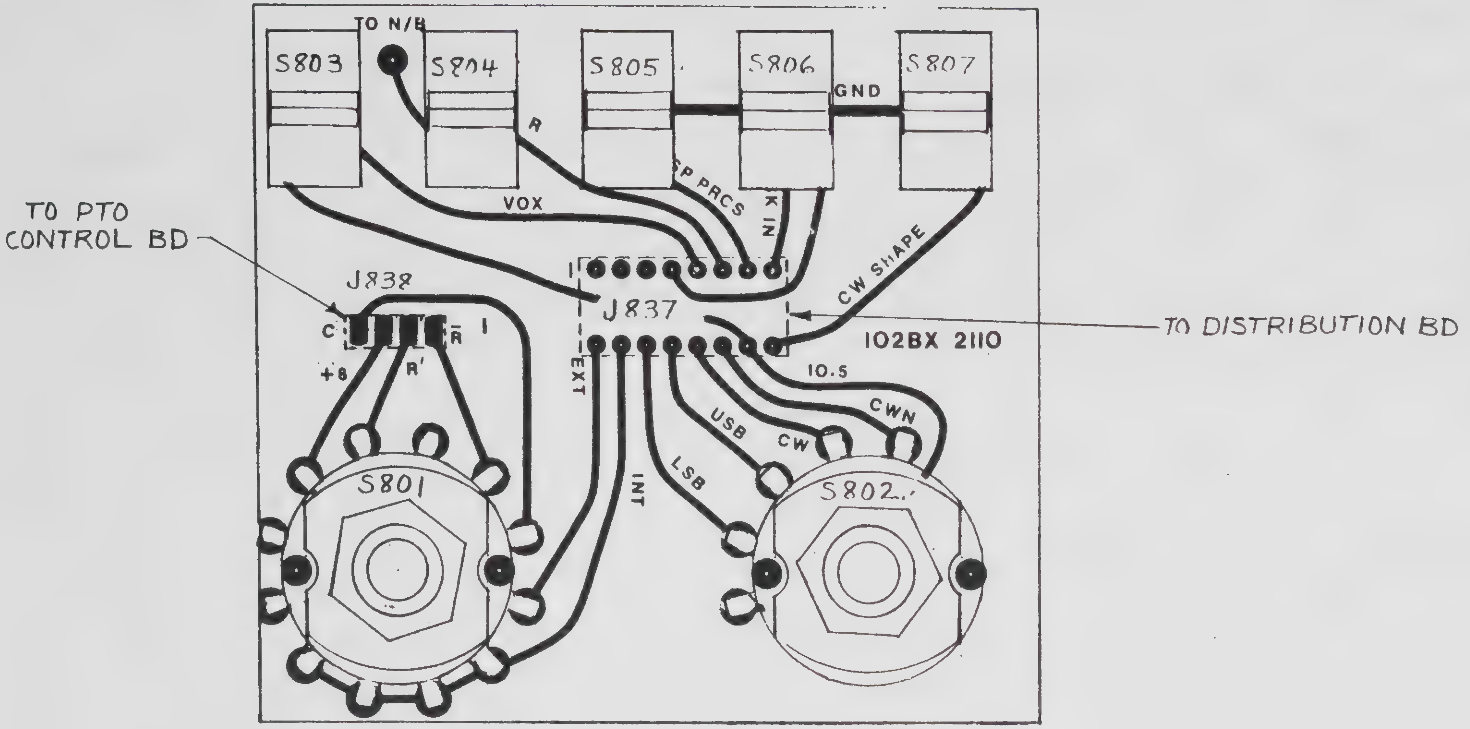


SWAN ELECTRONICS

SCALE NONE
APPROVED BY: JW
DRAWN BY C. DAVIDSON
REVISED
DATE 8-27-79

DISTRIBUTION BRD SCHEMATIC

DRAWING NUMBER
1028-2409



3. J837 & J838 ON FARSIDE.
2. COMPONENT DESIGNATOR SERIES 7800.
1. REF. ELECTRICAL ASSY 102BX-2010
REF. CIRCUIT BOARD 102BX-2110
REF. DRILL DRAWING 102BX-2210
NOTES:

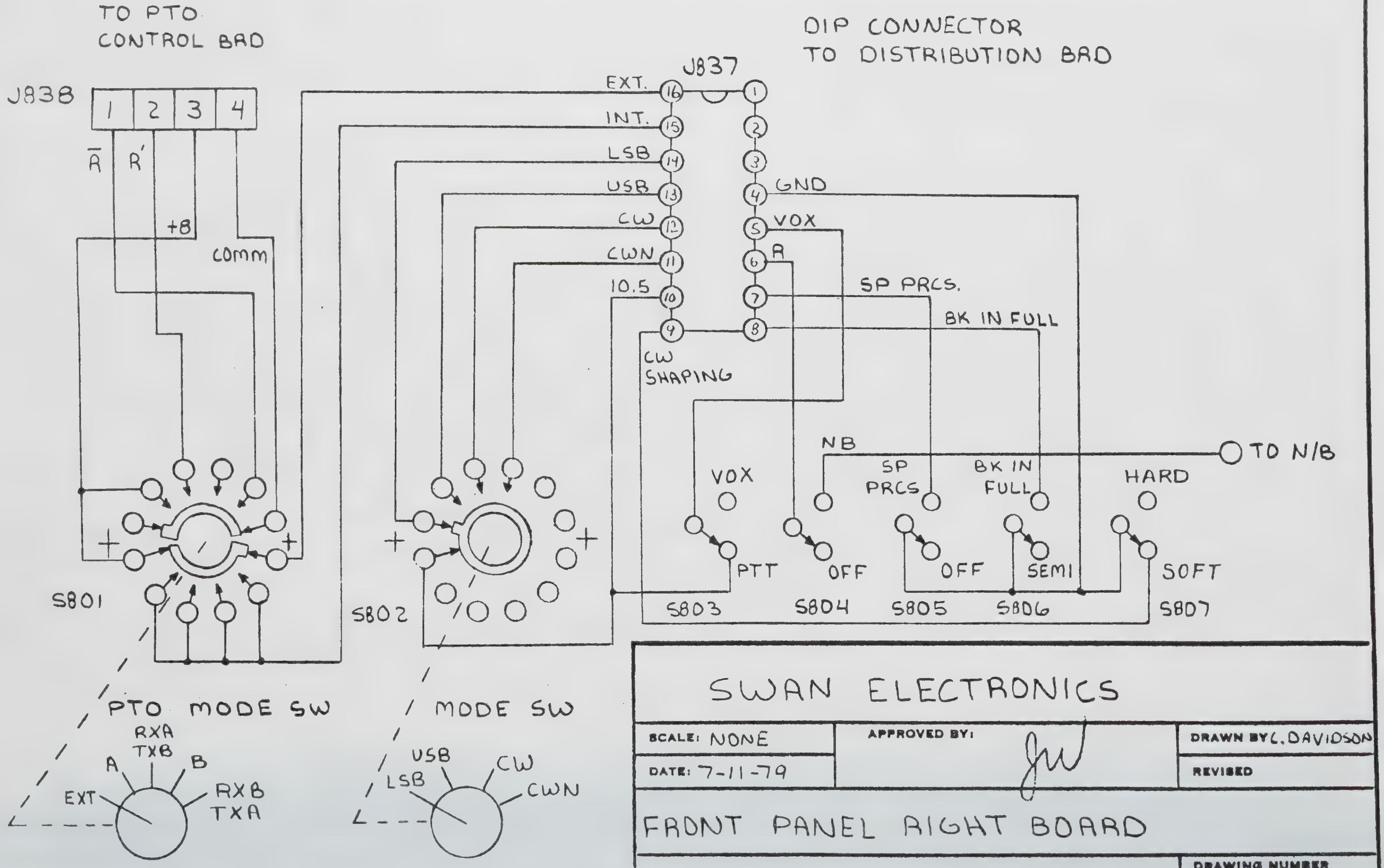
SWAN ELECTRONICS

SCALE: 2 : 1	APPROVED BY: Jw	DRAWN BY AJG
DATE: 9-12-79		REVISED

FRONT PANEL RIGHT BD LAYOUT

DRAWING NUMBER 102BX-2310

11X17 PRINTED ON NO. 1000H CLEARPRINT

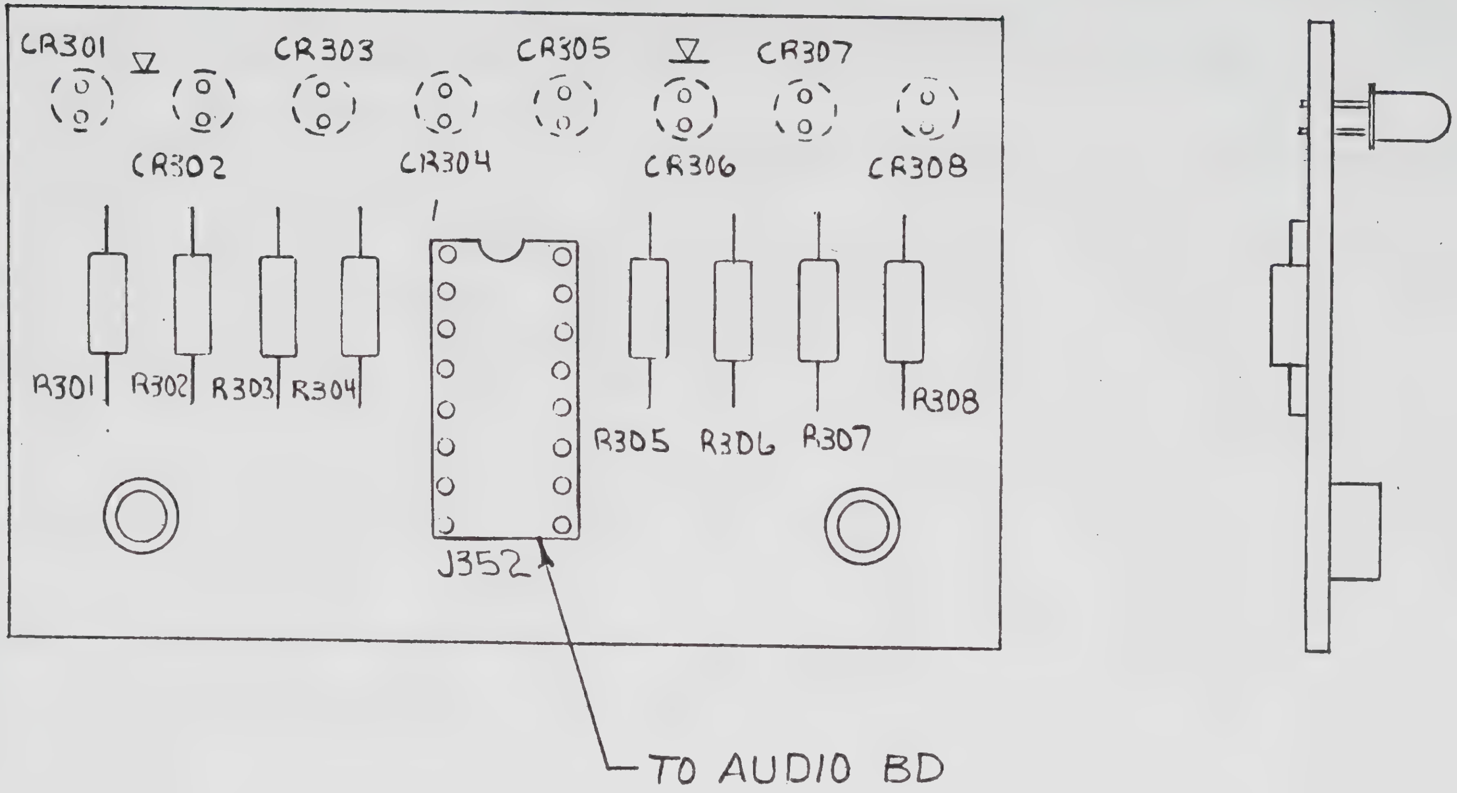


SWAN ELECTRONICS

SCALE: NONE
APPROVED BY: JW
DRAWN BY C. DAVIDSON
DATE: 7-11-79
REVISED

FRONT PANEL RIGHT BOARD

COMPONENT DESIGNATORS (7800 SERIES)
DRAWING NUMBER 102BX-2410



3. REF. ELECTRICAL ASSY 102BX-2011
   REF. CIRCUIT BOARD 102BX-2111
   REF. DRILL DWG 102BX-2211
2. LEDS MOUNTED FARSIDE.
1. COMPONENT DESIGNATORS 7300 SERIES

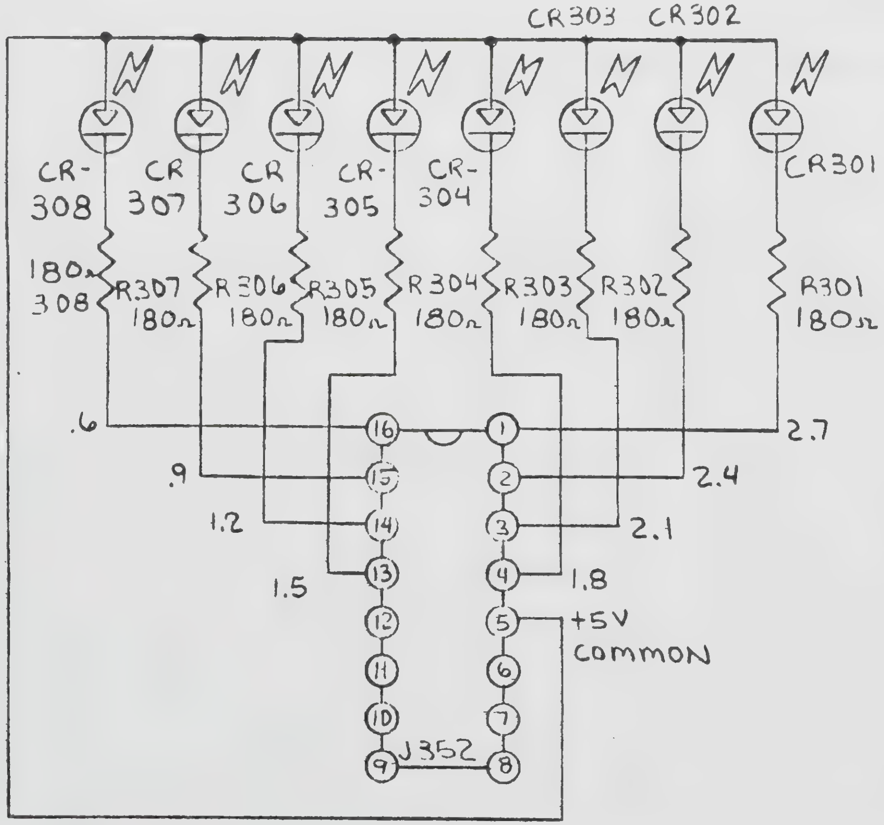
NOTES

SWAN ELECTRONICS

SCALE: NONE
DATE:
APPROVED BY:
DRAWN BY AJG
REVISED

PASSBAND IND. BD LAYOUT
102 BX

DRAWING NUMBER
102BX-2311



DIP CONNECTOR
FROM AUDIO BRD.

SWAN ELECTRONICS

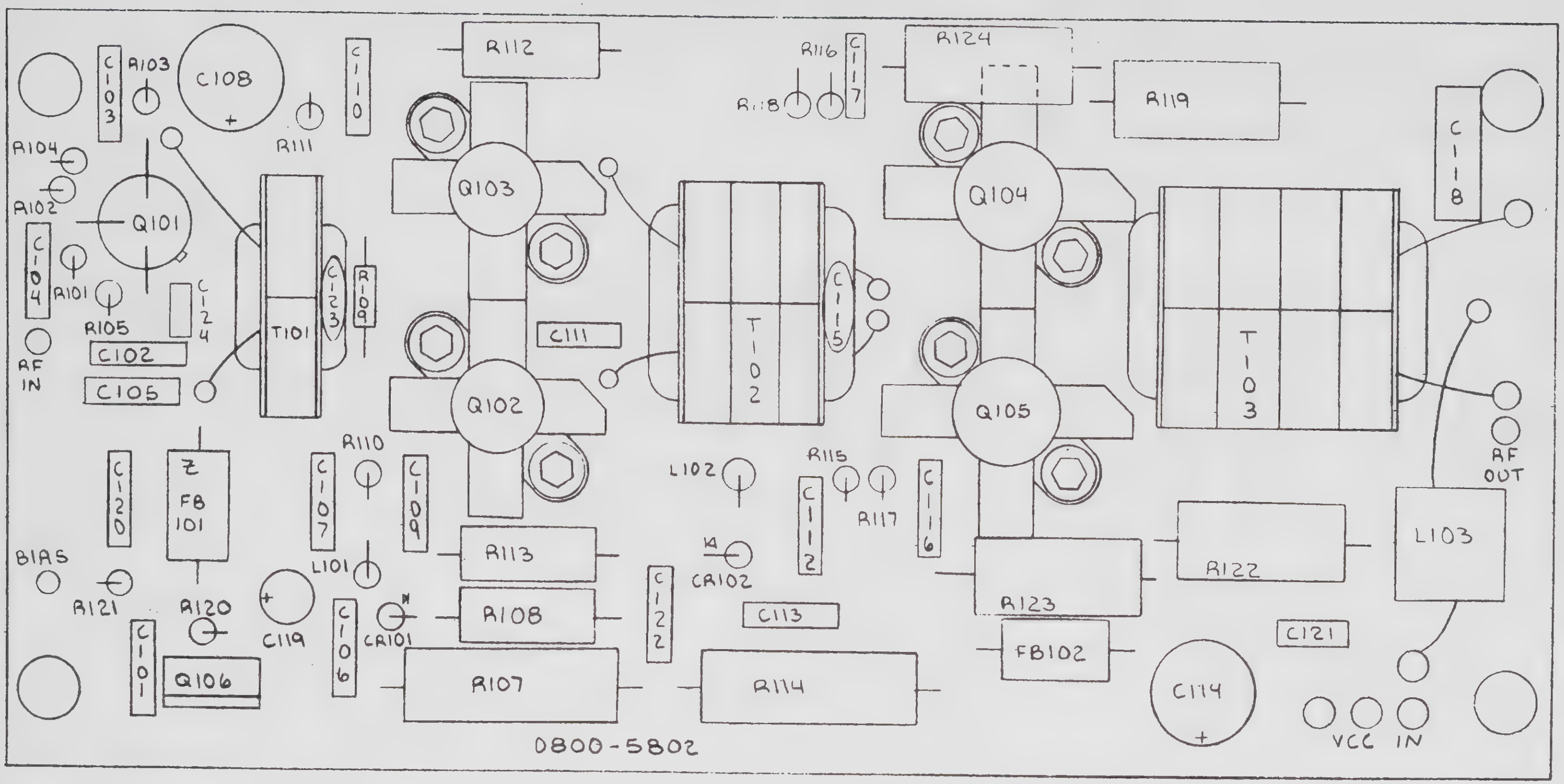
SCALE: NONE
DATE: 7-10-79
APPROVED BY:
DRAWN BY C. DAVIDSON
REVISED

PASSBAND INDICATOR BOARD SCHEMATIC

DRAWING NUMBER
102B-2411

1. COMPONENT DESIGNATORS 7300 SERIES

NOTES:



SWAN ELECTRONICS

SCALE: NONE

DATE: 1-2-78

APPROVED BY:

DRAWN BY: C.DAVIDSON

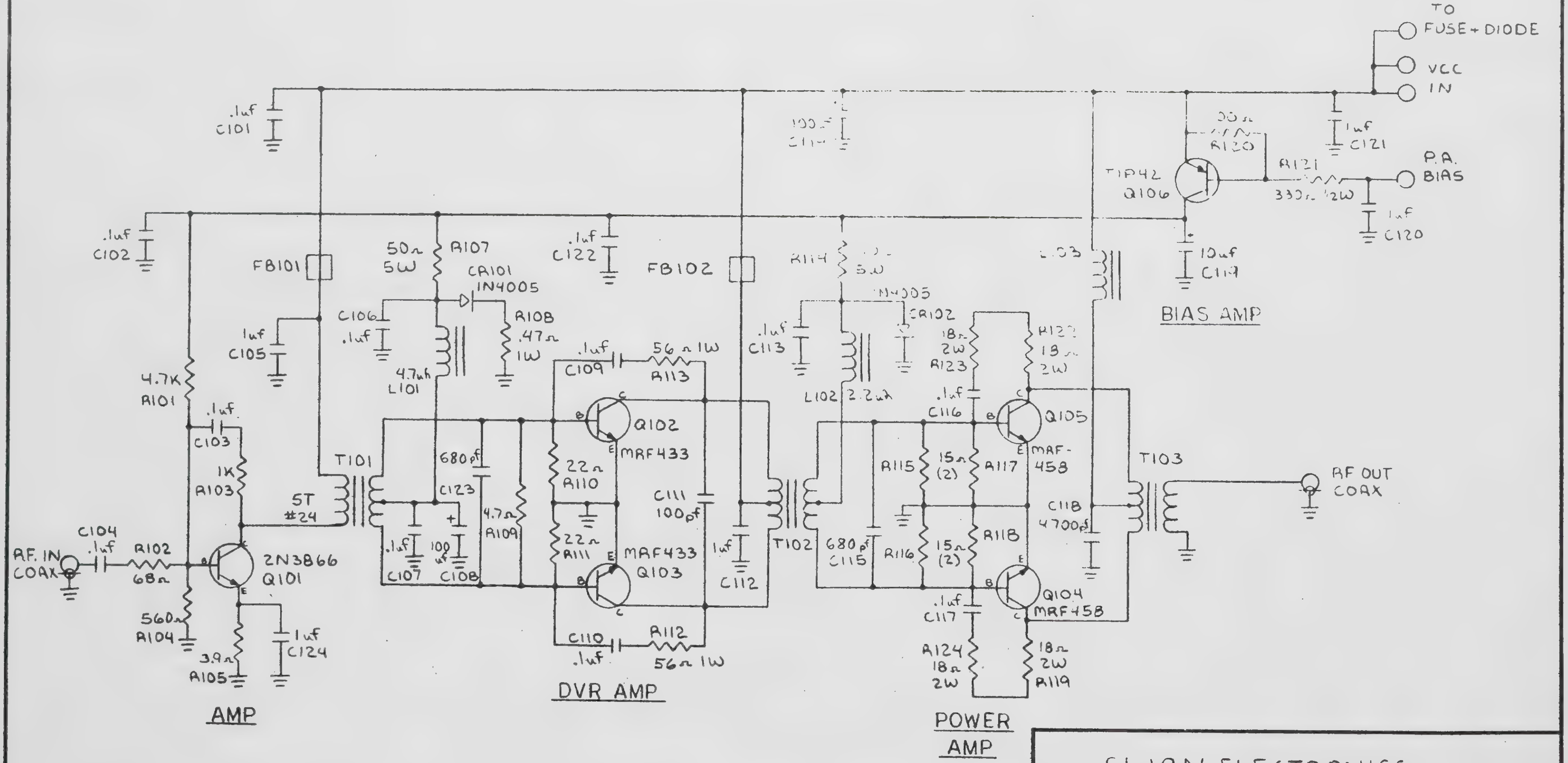
REVISED

POWER AMP BOARD LAYOUT

DRAWING NUMBER 0800-2302

1. COMPONENT DESIGNATORS (6000 SERIES)
NOTES:

11X17 PRINTED ON NO. 1000H CLEARPRINT



SWAN ELECTRONICS

SCALE: NONE	APPROVED BY:	DRAWN BY C DAVIDSON
DATE: 1-2-78		REVISED

POWER AMP BOARD SCHEMATIC

REV	DRAWING NUMBER
B	0800-2402B

LTR	DESCRIPTION	DATE	APP'D
ECN# B	REPLACED ALL .05uf CAPS WITH .1uf CD 5-30-79	6-1-79	JW

REVISIONS

NOTES

1. COMPONENT DESIGNATORS (6000 SERIES)

11X17 PRINTED ON NO. 1000H CLEARPRINT

SERVICE
MANUAL
ADDENDUM

ASTRO-103
TRANSCEIVER

MODIFICATION
TO THE
ASTRO-102BX
TRANSCEIVER

CUBIC COMMUNICATIONS, INC.
A member of the Cubic Corporation family of companies

Contents

Section One
General Information.........................1.1

Section Two
Specifications...............................2-1

Section Three
Receiver Theory of Operation.................3-1

Section Four
Transmitter Theory of Operation.............4-1

Section Five
Oscillator Theory of Operation..............5-1

Section Six
Ancillary Circuits Theory of Operation......6-1

Section Seven
Servicing Hints.............................7-1

Section Eight
Servicing Data..............................8-1

Section Nine
Parts Lists.................................9-1

CUBIC
COMMUNICATIONS
SWAN DIVISION

General Information

The Swan ASTRO 102BX is a 160M through 10M SSB amateur transceiver that features dual PTO's and frequency counter for split frequency operation with digital frequency readout. Frequency selection is the only tuning necessary when operating into a matched antenna with a VSWR of 1.7:1 or less.

Standard features include: Full break-in or semi-break-in CW operation; narrow band crystal CW filter; two position CW transmit pulse shaping; VOX; noise blanker; SSB transmit speech processor; true RF and IF gain controls; peak reading wattmeter with FWD, REF, and ALC positions; variable AGC decay; passband tuning; and audio notch filter. The basic single conversion design has been proven in thousands of the popular Swan transceivers. Mechanical and electrical designs are exceptional, for years of trouble-free operation, even in a rugged mobile environment.

The unit is designed for fixed and mobile use with minimum interconnections. For fixed station operation an auxiliary speaker/power supply, model PSU-6, provides 13.6 V DC from 115/230 V AC. A matching antenna tuner, model ST2A, is also available.

The transmitter may be used for RTTY or Slow Scan TV, but operation in these modes requires additional forced air cooling. See the operator's anual for precautions to be observed when operating in these modes. The ASTRO 102BX includes automatic gain control (AGC) in the receive mode, and automatic level control (ALC) in the transmit mode. CW sidetone is also provided.

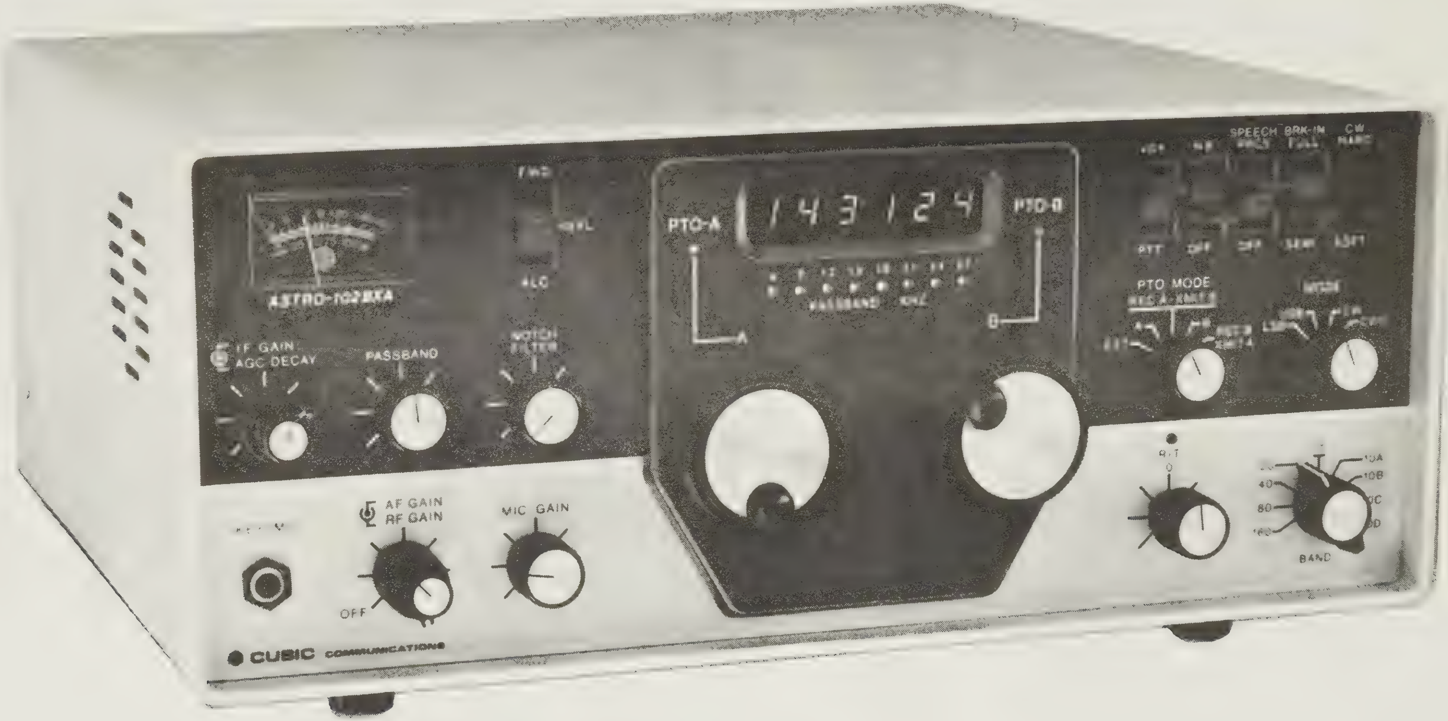

ASTRO-102BXA
FWD
REFL
ALC
IF GAIN
AGC DECAY
PASSBAND
NOTCH FILTER
PTO-A
143124
PTO-B
PASSBAND KHZ
SPEECH
BRK-IN
CW
PTT
OFF
PTO MODE
MODE
AF GAIN
RF GAIN
OFF
MIC GAIN
BAND
CUBIC COMMUNICATIONS

FIGURE 5-1

CUBIC
COMMUNICATIONS
SWAN DIVISION

Specifications

General

Frequency Coverage*

160M	1.8-2.0 MHz
80M	3.5-4.0 MHz
40M	7.0-7.5 MHz
20M	14.0-14.5 MHz
15M	21.0-21.5 MHz
10M	28.0-28.5 MHz
	28.5-29.0 MHz
	29.0-29.5 MHz
	29.5-29.999 MHz

*Approximately 50 to 100 kHz overrange provided on each band.

Frequency Control: Dual PTO with 6-digit frequency counter and RIT control.

PTO Modes of Operation: A only; Receive A-Transmit B; B only; Receive B-Transmit A; External LO.

Modes of Operation: SSB with selectable sideband. CW with automatic 800 Hz offset on transmit and full or semi-breakin.

Power Input Requirements: 12-14 V DC negative ground only. No damage up to 15 VDC. 20 amperes peak current.

Electromagnetic Interference:

Transmit: Five element low pass filter on each band.

Receive: Five element lowpass filter plus three pole bandpass filter on each band.

Dimensions: 6.375" H x 14.25" W x 13.25" D
16.2 cm x 36.2 cm x 33.7 cm
Weight: 23.5 pounds (10.6 kg)

Rear Panel:

Power connector:

Pin 1	13.6 V DC
2	Ground
3	Ground (used for accessories)
4	13.6 V DC
5	Ground
6	13.6 V DC fused to 3 amps (used for accessories)

Antenna connector: UHF (SO-239)

External speaker/phones: Phone jack

External relay: RCA jack

Key:	RCA jack
EXT. Modulation:	RCA jack
Fuse:	3 amp fast blow
Ext. L.O.:	RCA jack

Receiver

Circuit Design:	Single conversion to 9.00165 MHz IF using double balanced mixer. Exceptional immunity to overload and cross modulation.
Receiver Sensitivity:	10 dB S+N/N typical at 0.35 uV
Image Rejection:	Better than 60 dB
Receiver Selectivity:	SSB and CW--2.4 kHz bandwidth, two 8-pole crystal filters with shape factor 1.4; 6 dB to 100 dB CWN--300 Hz bandwidth. IF crystal filter in series with one 8 pole SSB filter
Passband Tuning:	SSB and CW eight pole cutoff continuously variable high pass or low pass. LED readout shows equivalent audio passband. CWN--IF crystal filter continuously tunable over 300 to 3000 Hz with passband control
Dynamic Range:	AGC greater than 100 dB; third order intercept +15 dBm
Audio Output Power:	Greater than 3 watts into 4 ohms.
Meter:	S-units from 1 to 9; 20, 40 and 60 dB over S-9

Transmitter

Circuit Design:	Excellent harmonic and TVI suppression. ALC limits SSB peak power and CW power to 100 watts output.
Power Input:	235 watts PEP input at 13.6 V DC input
Power Output:	100 watts PEP into 50 ohm non-reactive load at 13.6 V DC
Unwanted Sideband Rejection:	>60 dB down at 1000 Hz audio note
Harmonic Output:	>45 dB below peak power level
Carrier Suppression:	>50 dB below peak power level

Spurious Output:	> 55 dB below peak power level
Transmit Control: (for SSB)	PTT and VOX standard
CW Transmit:	Full or semi-breakin with sidetone standard. Automatic 800 Hz offset on transmit frequency.
Microphone Input:	47 kΩ input impedance
Audio Response:	300 to 3000 Hz ± 6 dB
Meter:	In receive mode, reads S units and dB over S-9; in transmit mode, reads peak power selected by meter switch: Forward power 100 watts full scale, reflected power 10 watts full scale, or reads ALC.
Cooling:	Large capacity heat sink fins supplied. For SSTV, RTTY, and semi-continuous transmit, forced air cooling such as a small "muffin" fan on heat sink fins is required.
VSWR Shutdown:	Full power up to VSWR = 1.7:1 Approximate limit ratio as follows:

VSWR	Percent Power
1.7:1	100%
2.0:1	80%
3.0:1	60%
Open or short	25% (equivalent voltage)

CUBIC
COMMUNICATIONS
SWAN DIVISION

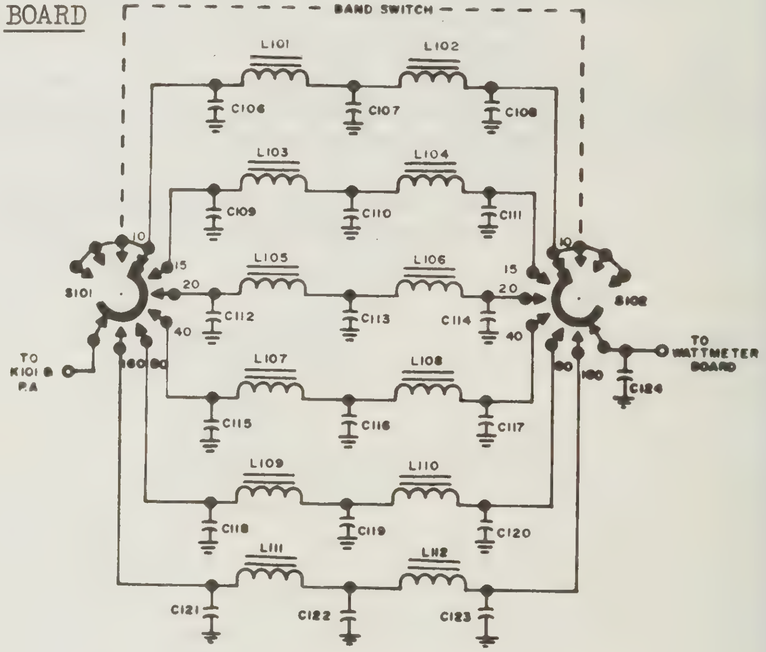
Receiver Theory

Bandswitch and Low Pass Filters................3-1
Preselector......................................3-2
First Mixer......................................3-3
Noise Blanker Gate...............................3-4
Filter, Buffer, and Post Filter Amplifier.......3-5
AGC Attenuator...................................3-5
Passband Tuning..................................3-6
CW Narrow Filter.................................3-7
Main I.F. Amplifier..............................3-8
Product Detector/Audio Mute......................3-8
Notch Filter.....................................3-9
Audio Power Amplifier/Receiver Mute.............3-9
Automatic Gain Control..........................3-10

NOTE: Component designations used in the "Theory" sections of this manual are the shortened numbers as they appear on the schematic diagrams. To determine complete component designation, the board/location number must be added after the letter or letters. See the "Note" in the first portion of the "Parts" section of this manual.

Schematic diagrams contained in the "Theory" sections of this manual are simplified to help in the understanding of the operation of the circuit; they may not always, therefore, be sufficient for troubleshooting purposes. When servicing the equipment, it is urged that the technician use only the schematic diagrams included in the "Servicing Data" portion of this manual.

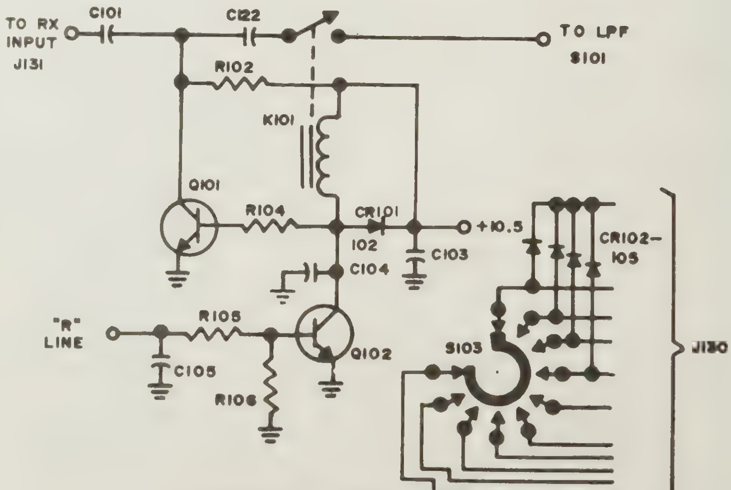
BANDSWITCH AND LOW PASS FILTER BOARD



The received signal from the antenna connector is passed through one of six low pass filters, which are also used in the transmit condition. Each is composed of two inductors and three capacitors, yielding a five pole filter for each band (160 through 10 meters). Note that the same filter is used for all segments of the 10 meter band.

The filters are selected by S101 and S102, sections of the BAND switch. From S101, the signal is passed to relay K101, through C122, and through C101. From there it is routed to the input of the preselector board.

Receiver antenna switching is accomplished by K101, which is activated by Q102. When the receive ("R") line is high, as it is in the receive condition Q102 is biased to saturation, energizing K101 and connecting C122 to the output of the selected low pass filter. Since the collector of Q102 is low when it is thus saturated, Q101 is in a non-conducting condition and has no effect. However, when the "R" line is low, as



RECEIVE ANTENNA SWITCHING

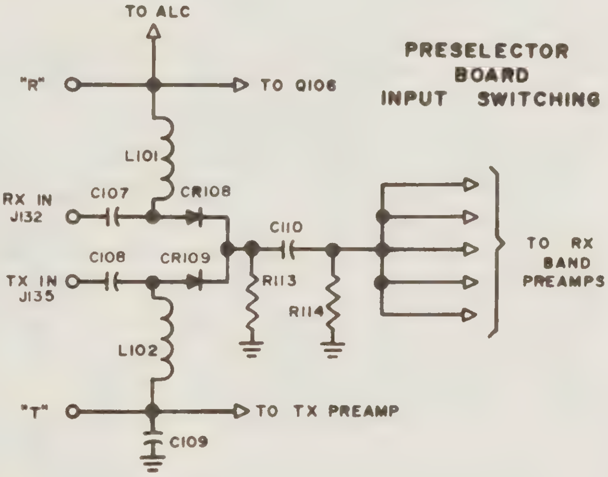
as it is in the transmit condition, Q102 is cut off, and its collector voltage is high. This biases Q101 on, providing a path to ground for any transmit signal coupled across the open contacts of K101 due to stray capacity. In this manner, the input of the receiver is protected during transmission periods.

S103, a part of the BAND switch, is used to provide a voltage to each of ten control lines, depending upon its position. There are separate control lines for the 160, 80, 40, 20, 15, 10A, 10B, 10C and 10D positions, and an additional line, which, through CR102 through CR105, is high for any of the 10 meter positions of the BAND switch. These control lines are used in other sections of the transceiver.

PRESELECTOR

In the receive condition, the transmit ("T") line is low. Thus, CR109 in the input circuitry of the preselector is reverse biased, and the TX IN jack is effectively open circuited. L102 is an RF choke to prevent signal from being bypassed to ground through C109 on the "T" line.

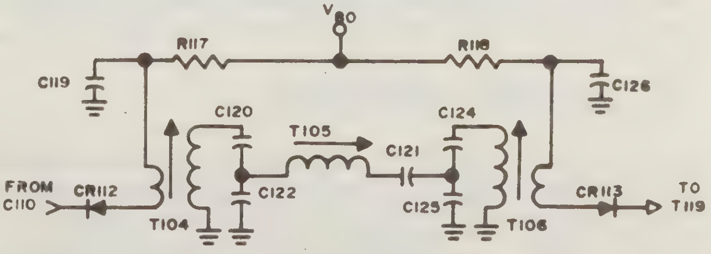
On the other hand, the "R" line is high in the receive condition, and is connected through RF choke L101 to CR108, which it biases to an "on" condition. R113 provides a ground return and the voltage drop across it reverse biases the "off" diode. C110 is a DC blocking capacitor, and passes the signal on to the input of the preselector band filters. R114 acts as a DC return for the switching currents for the filter selection circuits, and provides reverse bias to the "off" diodes.



PRESELECTOR BOARD INPUT SWITCHING

There are six band filters in the preselector (only one is used for all four segments of the 10 meter band). These filters are selected by the BAND switch, utilizing the voltages on the control lines from S103, as mentioned earlier. For discussion purposes, only the 80 meter filter will be considered.

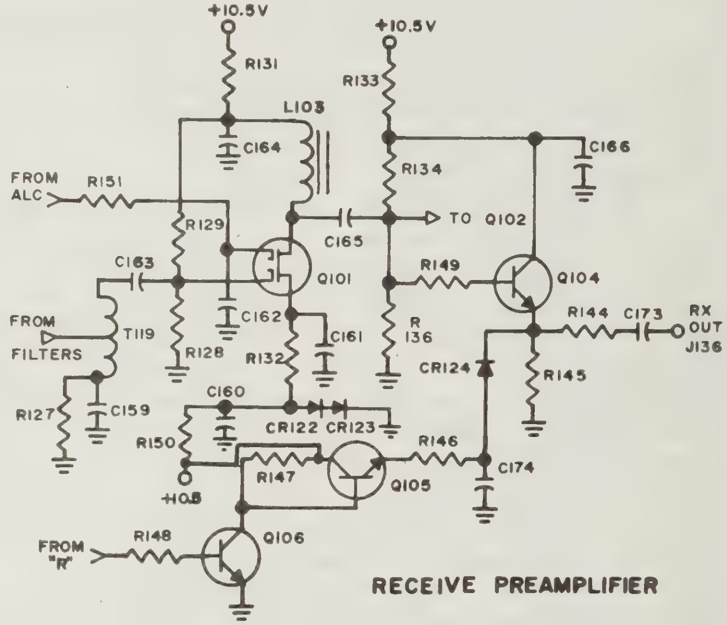
The voltage from the 80 meter control line is coupled through isolating resistor R117, the primary of T104, and CR112 to the output of the input switching circuit previously discussed, where it is provided with a ground return path through R114. C119 provides a ground for the signal at one end of the primary of T104. The signal is then passed through diode switch CR112 to T104, while the voltage drop across R-114 reverse biases the corresponding diode switches in the other filters.



80 M BAND FILTER

The secondary of T104 is resonated by the C120-C122 series combination, and they act as an impedance transformation voltage divider. T105 and C121 form a series resonant circuit, and its output is coupled to C124-C125, which again transforms the impedance of the signal path, and also resonates with output transformer T106. The combination of the two parallel resonant circuits and one series resonant circuit forms a bandpass filter of excellent characteristics. Output switching is provided by R118, in conjunction with CR113, in a manner identical to the input switching scheme.

The output of the preselector filters is coupled to the preamplifier stage, Q101, via autotransformer T119 and coupling capacitor C163. R127 provides a DC path to ground for the switching diode, and provides reverse bias for the "off" diodes in the other filters, while C159 effectively grounds the bottom of T119 for the signal.



RECEIVE PREAMPLIFIER

Bias for Q101 is provided both by the source circuitry and the R128-R129 combination. So far as gate two is concerned, the bias is determined by the source network as well as whatever is applied to it via the ALC line. CR122 and CR123, biased in the forward direction by the voltage applied by R150, provide a fixed drop of approximately 1.4 volts, while additional bias is supplied by the drop across source resistor R132. C160 and C161 are both bypasses.

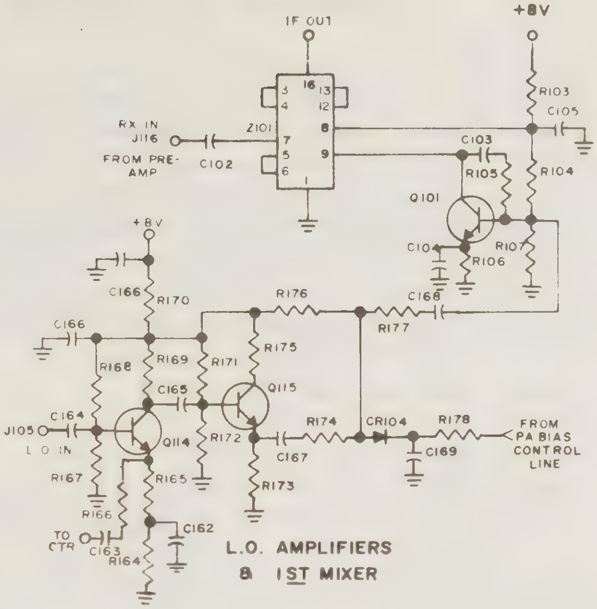
Signal from the drain of Q101 is taken via C165, after being developed across load L103. R131 and C164 form a decoupling network for the supply line.

In the receive condition, the "R" line is high, causing Q106 to saturate, thus lowering the voltage on the base of Q105, causing it to cease conduction. Diode switch CR124 is then reverse biased by the voltage drop across R145, allowing Q104 to operate in a normal manner. Bias for Q104 is derived by R134 and R136, while R149 serves to prevent instability. Output of the preselector is taken from the emitter of Q104 via R144 and C173 and coupled to the receiver input of the exciter board.

EXCITER BOARD--FIRST MIXER

After passing through the preselector and being amplified by the preamplifier, the received signal is connected to mixer Z101 via C102, a blocking capacitor. Local oscillator signal, brought to the exciter board through J105, is coupled via blocking capacitor C164 to the base of amplifier Q114. Bias for Q114 is developed in a conventional manner via the R167-R168 voltage divider. The emitter resistor of Q114 is divided into two parts, R164 and R165. While the former is bypassed in the usual manner, the latter is not, and a signal suitable for driving the display frequency counter is taken from across it, via R166 and C163.

Amplified L.O. output is taken from the collector of Q114 and coupled to the base of Q115. Again, bias is conventional, being developed by R171 and R172. The output of Q115 is taken across its emitter resistor, R173, via C167 and R174, and coupled through R177 and C168 to the base of mixer driver Q101. The P.A. Bias control line is high in the receive condition, thus reverse biasing CR104, preventing the L.O. signal from being shunted to ground by C169, as it would be in the transmit condition.

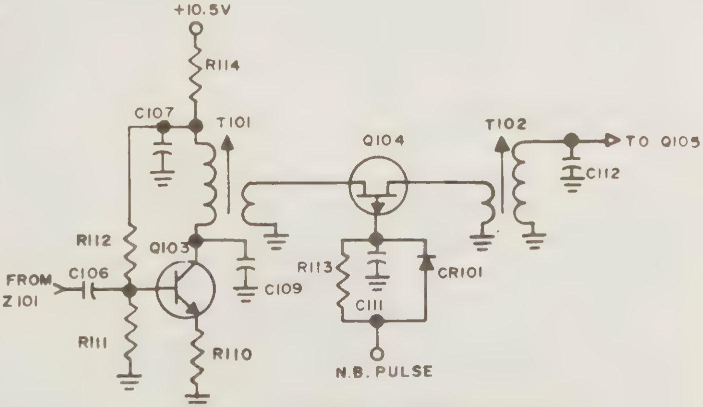


L.O. AMPLIFIERS & 1ST MIXER

Bias for Q101 is developed by R104 and R107, and C104-R106 form the usual emitter circuit. C103 and R105 form a feedback network, which, in conjunction with R177 and R174 set the gain of Q101. For a more complete discussion of feedback amplifiers and their operation, refer to the "Servicing Hints" section of this manual.

NOISE BLANKER GATE

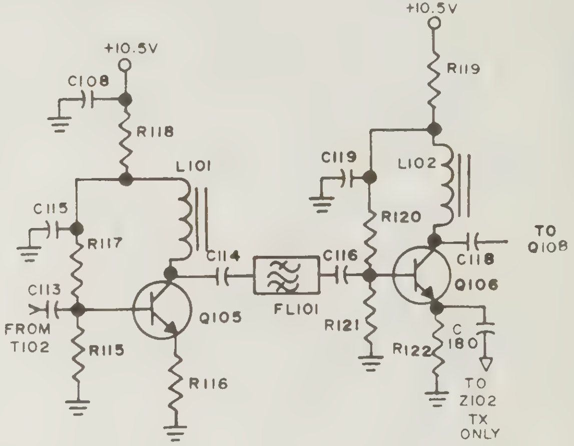
The output of the first mixer, Z101, now at the I.F. frequency, is coupled to the first I.F. amplifier, Q103, via blocking capacitor C106. R111 and R112 form the bias network, while emitter resistor R110 is unbypassed, providing some degenerative feedback. Output is developed across a parallel-tuned resonant circuit composed of the primary of T101 and C109. R114 and C107 form a decoupling network, while the latter also places the upper end of the primary of T101 at AC ground, so that C109 is effectively in parallel with T101.



FIRST I.F. AMPLIFIER & NOISE BLANKING GATE

The secondary of T101 presents a signal at low impedance, which is coupled through noise blanker gate Q104 to the primary of T102, where it is stepped back up. C112 resonates the secondary of T102. When the noise blanker circuit detects a noise pulse of sufficient amplitude, the gate of Q-104 is fed a positive pulse through CR101, effectively cutting it off and opening its source-to-drain path, thus opening the signal path. Otherwise, a ground is applied to the gate via R113, keeping Q104 in an "on" condition. C111 and R113 serve to lengthen the pulse at the gate of Q104 to ensure positive blanking.

FILTER BUFFER, FILTER, AND POST FILTER AMPLIFIER



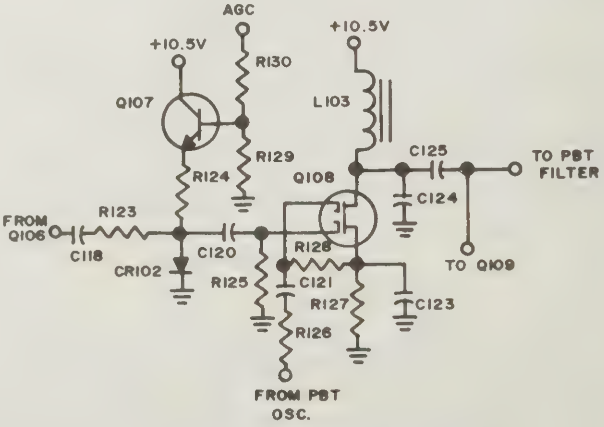
FILTER BUFFER, FILTER, AND POST FILTER AMPLIFIER

From T102, the signal is coupled to filter buffer amplifier Q105 via blocking capacitor C113. Bias is developed by R115 and R117, while emitter resistor R116 is unbypassed, providing degeneration. Output is taken across collector load L101 and coupled through blocking capacitor C114 to the first I.F. filter, FL101. R118, C108, and C115 are a decoupling network.

From filter FL101, the signal is coupled through C-116 to the base of post filter amplifier Q106, which is biased by R120 and R121. In the transmit condition, Q106 acts as an emitter follower (discussed in detail in the Transmit section of this manual), while in the receive condition, the output is taken from the collector after being developed across load L102. C118 is a blocking capacitor. Note that the use of inductive loads and unbypassed emitters in the stages preceeding the AGC attenuator ensure good strong signal handling capability.

AGC ATTENUATOR

From Q106, the received I.F. signal is coupled through C118 and R123 to AGC attenuator PIN diode CR102, and thence via C120 to gate one of pass band tuning first mixer, Q108. AGC voltage, when required, is applied to the base of AGC driver Q107 via voltage divider R129-R130. Current is then supplied to the PIN diode via R124, causing it to conduct to a degree determined by the AGC voltage; it acts, in conjunction with R123, as a variable voltage divider, similar in effect to a conventional volume control, and thus provides the AGC action required. Up to about 50 dB of attenuation is available at this point.

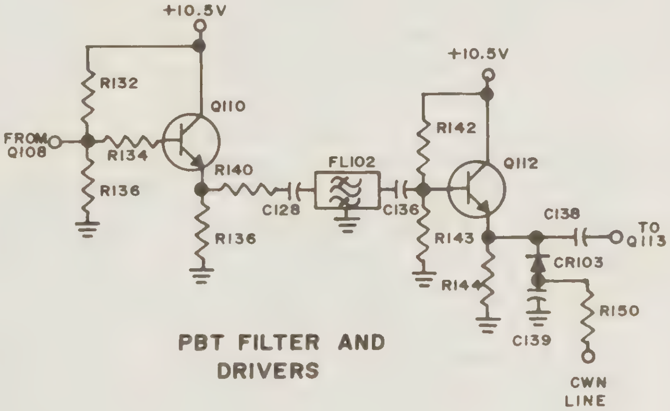


I.F. AGC ATTENUATOR AND
PBT 1st MIXER

PASSBAND TUNING

Q108, a dual-gate MOSFET, acts as a subtractive mixer, with the first I.F. signal applied to gate one, and the passband tuning oscillator signal applied to gate two. Bias for gate one is developed across R127, while connecting gate two to the source via R128 makes it operate without bias. Output of Q108, now at the pass band tuning I.F. of 13.8 MHz, is taken from its drain. L103 and C124 form a parallel resonant drain load. C125 is a coupling capacitor, routing the signal to the SSB filter via Q110 and to the CW narrow filter via Q109.

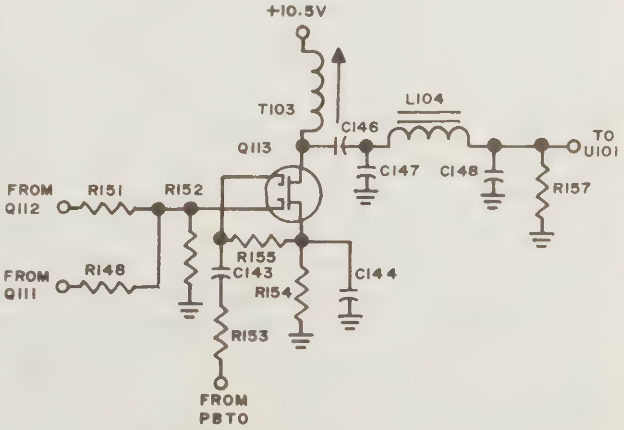
Q110, conventionally biased, acts as an emitter follower to drive the SSB filter, FL102. After passing through the filter, the signal is coupled to Q112 via C136. Q112 is also an emitter follower, with its output developed across R144 and bias developed by the R142-R143 combination. C138 is a blocking capacitor.



PBT FILTER AND DRIVERS

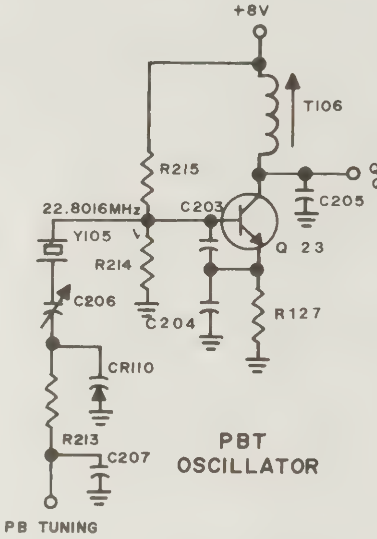
When operating in the CW narrow mode, voltage is applied to the CWN line through R150 to diode switch CR103. This voltage reverse biases the base-emitter junction of Q112, causing it to cease to conduct, and thus breaking the signal path. Also, C139 shorts the signal path through CR103.

Signal from the CWN filter amplifier Q111 (covered later) and from the SSB filter are coupled to gate one of the second PBT mixer Q113. R148 and R151 are isolation resistors, and R152 is the gate return resistance for Q113, which is also a subtractive mixer. Oscillator energy is coupled to gate two of Q113 in the same manner as it was to the first PBT mixer, and the bias arrangement is also the same. Output is taken via C146, now back at the first I.F. frequency of 9 MHz. C147 serves the dual function of resonating with T103 and serving as the first element in a low pass filter composed of C147, L104 and C148, which filters out unwanted mixer products. R157 terminates the filter in its required load.



2nd PBT MIXER

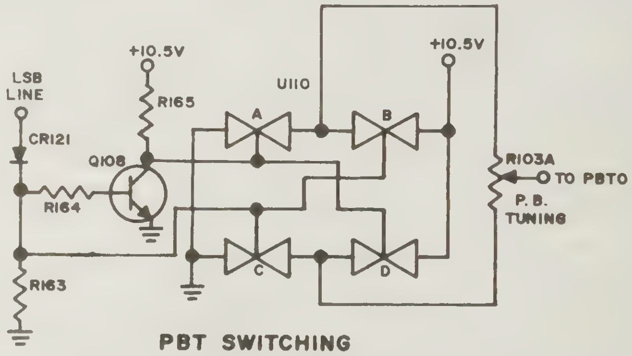
PBT oscillator Q123 is a conventional Colpitts crystal controlled oscillator, with C203 and C204 providing the feedback capacitive divider. R214 and R215 form the bias network, while R127 is the emitter resistance. T106 and C205 form a parallel resonant collector load. The frequency of the oscillator is adjusted by the application of a voltage from the PASSBAND tuning control on the front panel through isolation resistor R213, to varicap CR110. The range of adjustment is set by trimmer C206, while C207 is a bypass.



PBT OSCILLATOR

Because the same frequency oscillator signal is applied to both PBT mixers, any signal going into the first PBT mixer will leave the second on precisely the same frequency. This is why the PASSBAND tuning control only affects the relative position of the passband and has no effect on the frequency of the signal tuned in.

The direction of tuning of the passband oscillator must be one way for USB/CW operation, and the opposite direction for LSB operation for the effect at audio to remain the same. When in the LSB mode, a positive voltage is applied through CR121 on the audio board to the base of Q108, through limiting resistor R164. Simultaneously, it is also used to turn sections B and C of switch U110 on. Section B of U110 applies +10.5 volts to the upper end of the PASSBAND tuning control R103A, while section C grounds the other end. Since Q108 is saturated, its collector is low, and this voltage, applied to sections A and D of U110, turns them off. When in the USB or CW modes, however, the LSB line is low, and thus sections A and D of U110 are on, while sections B and C are off. This action reverses the voltage applied to the tuning control.

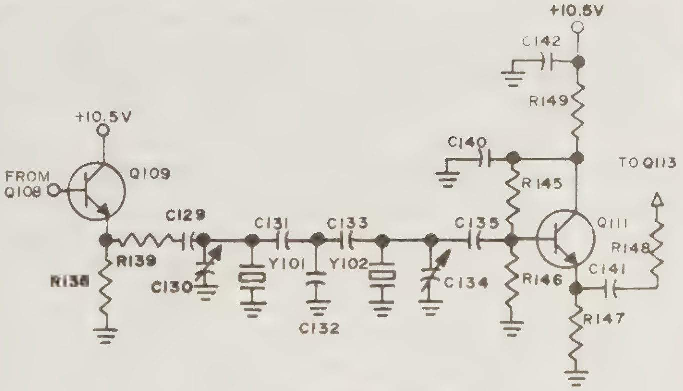


PBT SWITCHING

Operation of the PASSBAND tuning indicator will be covered in the "Ancillary Circuits" section of this manual.

CWN FILTER

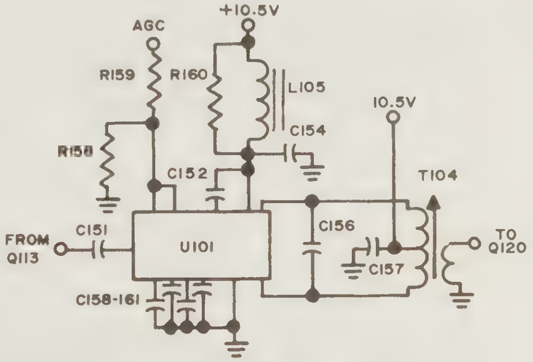
Signal at the second I.F. of 13.8 MHz, from the first passband tuning mixer (Q108), is also coupled to the base of Q109. Bias is received from the same network that supplies Q110, the filter driver. Q109 acts as an emitter follower driver for the CW narrow filter, and the required filter termination impedance is provided by R138 and R139, while C129 is a blocking capacitor. The CW narrow filter is composed of Y101 and Y102, with C131 and C133 acting as coupling capacitors and C130 and C134 providing adjustments to properly align the filter response. C135 acts as an output coupling capacitor, and carries the signal to the base of Q111. Power for Q111 is taken from the CWN line, and bias is provided by R145 and R146. Output is taken across emitter resistor R147 and coupled to the second passband tuning mixer Q113 through C141 and R148.



NARROW CW FILTER

MAIN I.F. AMPLIFIER

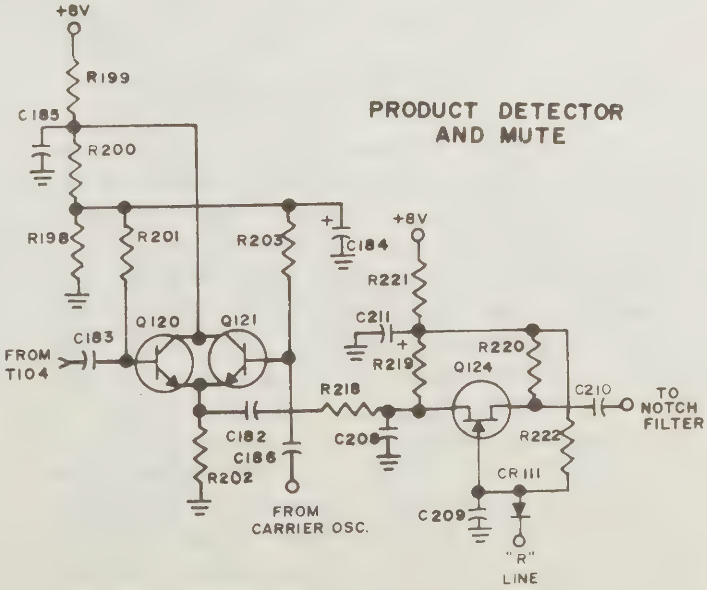
The signal from Q113, now again at the 9 MHz I.F., is coupled through C151 to the main I.F. amplifier, U101. This integrated circuit is composed of two amplifier sections, with the output of the first section developed across the R160-L105 load combination. C154 resonates with L105, while R160 broadens the response by "swamping" the tuned circuit. C152 couples the output thus developed from the first amplifier section of U101 to the second section of the same IC. Output of the second section is coupled to push-pull T104, which is resonated by C156. C157 is a bypass. The output of T104 is coupled to the product detector, Q120. AGC control is applied through voltage divider R158—R159. Overall gain of U101 is up to 70 dB.



MAIN I.F. AMPLIFIER

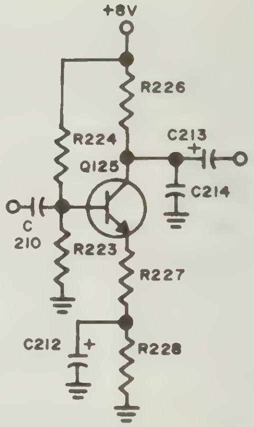
PRODUCT DETECTOR/AUDIO MUTE

Carrier oscillator energy is coupled via C186 to the base of Q121, which forms an emitter-coupled pair with Q120. Bias is developed by R200 and R198, acting as a voltage divider, and R201 and R203 serve as isolating resistors, while C184 is a bypass. Q120 and Q121 act as a mixer, combining the I.F. input from T104 and the carrier oscillator signals to recover the audio frequency of the incoming signal. Output is coupled via C182 to an RC network composed of R218 and C208,



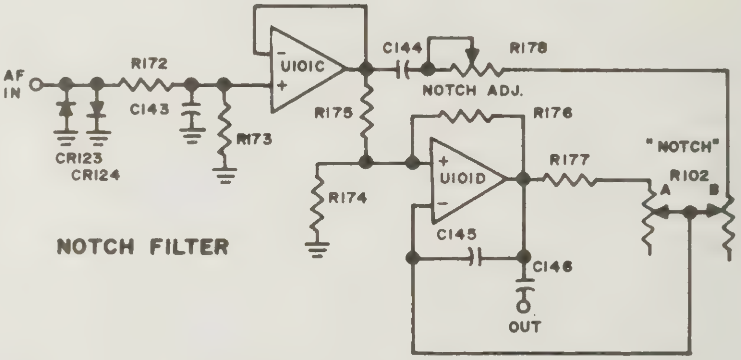
PRODUCT DETECTOR AND MUTE

which filters out any undesired mixer products. Q124, in series with the signal path, acts as a switch to cut off the receiver output AF when in the transmit condition. In that instance, the "R" line would be low, grounding the gate of Q124 and causing it to act as an open circuit. R219 and R220 provide bias voltage for the source and gate. Output of the AF mute gate is then coupled via C210 to the base of AF amplifier Q125, which provides additional gain. Bias is conventional, while the emitter resistance is only partially bypassed to set the gain (see the "Servicing Hints" section of this manual for a discussion of feedback amplifiers such as this). Output is taken via C213, while C214 provides a degree of high frequency rolloff for more pleasing audio response.



A.F. PREAMP

NOTCH FILTER



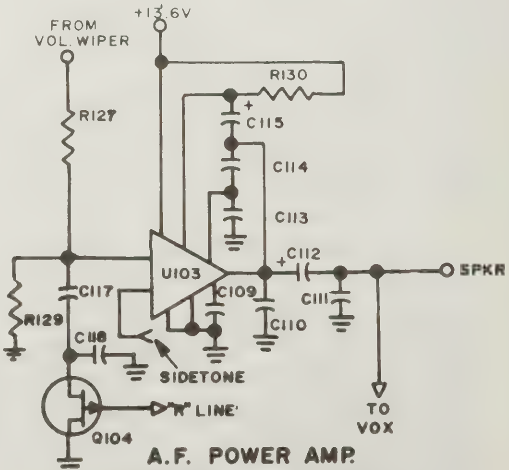
NOTCH FILTER

Audio from Q125 is applied to CR123 and CR124, which act as a limiter, effectively clipping off any peaks exceeding approximately ± 0.7 volt. The R172-C143 combination filters the output of the clipper and shapes the audio response. U101C is connected as a voltage follower, providing drive for active filter U101D that is the same amplitude and phase as its input, but of a lower driving impedance.

C144 and R178, with R120A on the front panel, form the series leg of a Wein bridge, while C145, R177 and R102B (also on the front panel) form the shunt leg of the same Wein bridge. This bridge is inserted in the inverting feedback loop to form a notch when combined with the normal audio frequency applied to the non-inverting input of amplifier U101D. R176 and R174 form a positive feedback path to raise the"Q" of the circuit.

AUDIO POWER AMPLIFIER/RECEIVER MUTE

Audio from the output of the notch filter is applied to the high side of the front panel volume control, and its wiper is returned, through R127, to the input of audio power amplifier U103. R129 acts in conjunction with R127 as a voltage divider. Receiver mute switch Q104 is connected between this input and ground, through isolating capacitor C117. When in the receive condition, a voltage on the "R" line holds Q104 in the "open" condition, allowing the signal to be amplified by U103; when in the transmit condition, the "R" line is grounded, turning Q104 "on" and shunting the signal to gr-



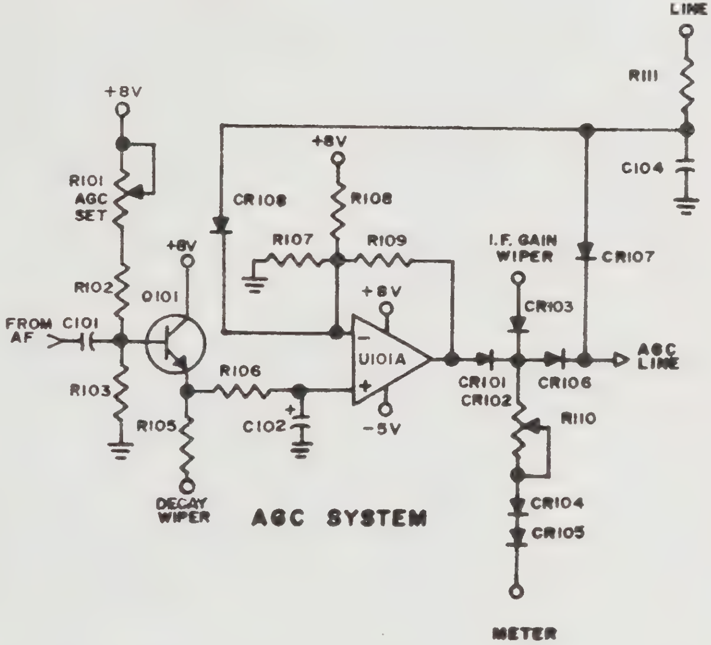
A.F. POWER AMP.
AND RECEIVER MUTE

ound. C118 and R127 shape the audio response.

When in the transmit condition, and when sending CW, the sidetone audio signal is applied to the other input of U103; thus, the operator is able to monitor his own sending.

C113 and C114 form an audio shaping network for more pleasing response, and C112 couples the output to the speaker. C110 and C111 are RF bypasses. C109 bypasses certain internal circuitry of U103, while R130 and C115 provide a type of feedback path known as "bootstrap".

AUTOMATIC GAIN CONTROL



The audio frequency signal is coupled to the base of Q101 via C101. Bias is set by R101, R102 and R103, with R101 serving as the AGC threshold adjustment. When the audio peaks are of sufficient amplitude, Q101 conducts, which causes a current to flow through R105 and the DECAY control, a variable resistance to ground. The voltage drop across this combination charges capacitor C102, with series resistor R106 controlling the charging time, and thereby the attack time. When the AF signal disappears, C-102 discharges through R106 and the R105-DECAY control combination, thus the setting of the control varies the decay time of the AGC.

The voltage across C102 is applied to the non-inverting input of amplifier U101A. Provided the output of U101A is greater than any reverse bias to CR101 supplied through CR103 by the I.F. GAIN control, the output of U-101A is applied to the AGC line through CR106, and to the S meter through calibration control R110 and voltage dropping diodes CR104 and CR105. R107, R108 and R109 establish a reference level and the gain of U101A.

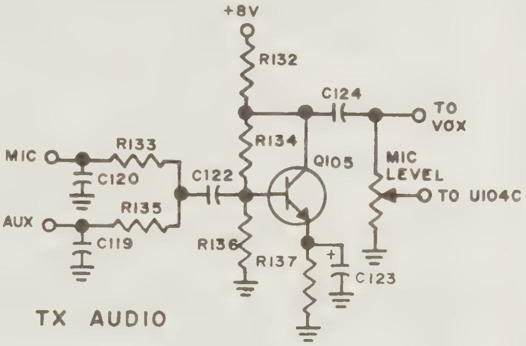
When in the transmit condition, the "T" line goes high, and applies a voltage through CR108 to the inverting input of U101A, causing its output to go low and remain there so long as the "T" line remains high, thus disabling the amplifier. This ensures that the S meter is disabled in transmit. This same voltage is applied, through CR107, to the AGC line to force maximum attenuation of the AGC PIN diode in the IF amplifier chain, and to cut off the I.F. integrated circuit amplifier, U101, on the exciter board. This ensures that no signal is present in the AGC loop during transmit and full gain is available when switched back to receive.

Transmitter Theory

Microphone Preamplifier.................4-1
Speech Processor........................4-1
Modulation Amplifier....................4-1
Balanced Modulator......................4-2
First I.F. Amplifier....................4-2
Noise Blanker Gate......................4-2
Filter Buffer, Filter...................4-3
Transmit Mixer..........................4-3
Preselector
   Input switching......................4-4
   Bandpass Filters.....................4-4
   Amplifiers...........................4-5
Power Amplifier
   Predriver............................4-6
   Driver Amplifiers....................4-6
   Output Amplifiers....................4-7
   P.A. Bias Control....................4-7
Low Pass Filters........................4-7

MICROPHONE PREAMPLIFIER

The audio signal from the microphone or external input jack is coupled through R133 or R135, respectively, to blocking capacitor C122 and thence to the base of microphone preamplifier Q105. C119 and C120 are RF bypasses, while bias is developed from the collector by voltage divider R134-R136. This provides feedback to keep the gain of the stage fixed. The emitter circuit is standard, composed of resistor R137 and bypass capacitor C123. Output is coupled via C124 to the VOX circuit and to the internal microphone level control.

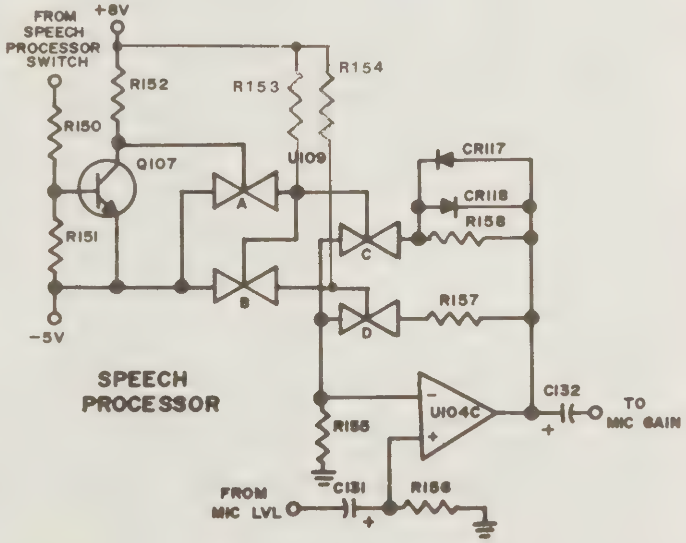


TX AUDIO

SPEECH PROCESSOR

When the operator desires to use the built-in speech processor, a voltage is applied, through voltage divider R150-R151, that biases Q107 into conduction thus lowering its collector voltage and turning section A of switch U109 off. This in turn activates sections B and C of that switching IC. The former turns section D off, opening the feedback path for U104C composed of R157, and the latter completes the alternative feedback path composed of R158, CR117 and CR118.

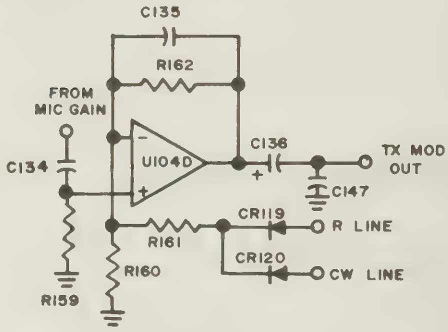
U104C, in this case, operates as a compressor. When signal peaks at the output of the amplifier exceed the barrier potential of CR117 and CR118 (± 0.7 volt) these peaks are fed back to the inverting input of that amplifier, thus reducing its gain. When speech processor operation is not desired, R157 is utilized as the feedback path, converting U104C into an ordinary fixed-gain amplifier stage.



SPEECH PROCESSOR

MODULATION AMPLIFIER

U104D is a fixed-gain amplifier, immediately following the front panel MIC GAIN control. Feedback is provided by the R162-C135 parallel combination, with C135 providing additional feedback at the higher audio frequencies thus reducing gain of the stage in that range, and providing a more pleasant normal sound to the transmitted voice signal. The output is taken via C136, with C147 providing bypassing for RF energy. When operating in the CW mode, or in the receive condition, the amplifier is disabled by voltages applied through CR120 or CR119, respectively, and through R161, to the inverting input of the amplifier, causing its output to go to its lowest limit and remain there so long as these voltages are present, preventing it from amplifying the transmit audio signal.

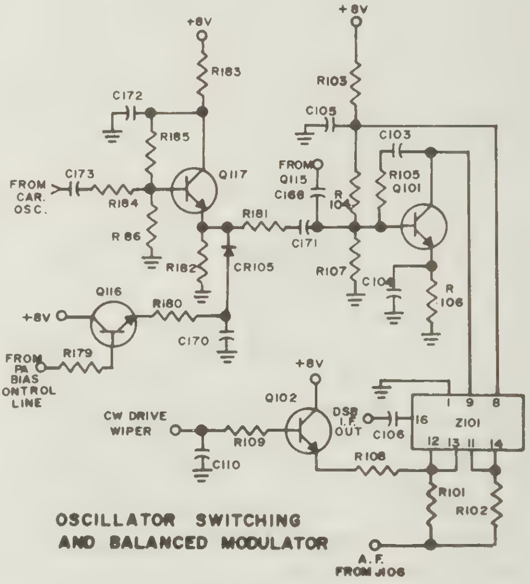


MODULATION AMPLIFIER

BALANCED MODULATOR

Z101, on the exciter board, used as the first mixer in the receive condition, is used as the balanced modulator in the transmit condition. Audio from the modulation amplifier is applied to Z101 through R101 and R102. Meanwhile, the P.A. bias line, which goes low for transmit, cuts off Q116, removing the voltage applied through R180 and CR105 from the emitter of Q117 which allows it to operate as a normal amplifier. This voltage, present in the receive condition, reverse biases the base-emitter junction of Q117, disabling it when it is not needed.

Carrier oscillator energy is coupled through C173 and R184 to the base of Q117, where it is amplified. Output is taken from the emitter, through R181 and C171, and applied to the base of Q101. (In the receive condition, local oscillator energy is coupled through Q115 to the base of Q101, but this path is interrupted in the transmit condition). Q101 is a feedback amplifier, with its gain determined by the ratio of feedback resistor R105 to the input series resistor, R181. Bias and emitter circuits are conventional, and output is taken from the collector of Q101 through a winding internal to Z101. The DSB suppressed-carrier output is taken through C106. (For internal circuitry of Z101, see the Servicing Hints section of this manual.)

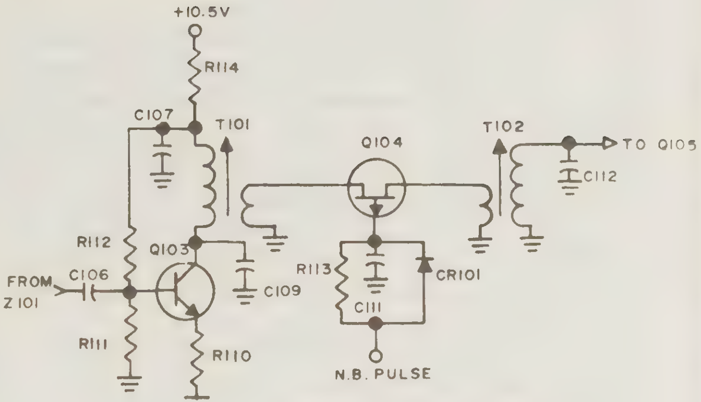


OSCILLATOR SWITCHING AND BALANCED MODULATOR

FIRST I.F./NOISE BLANKER GATE

The DSBSC output of Z101 is coupled to the first I.F. amplifier, Q103, via C106. R111 and R112 form the bias network, while emitter resistor R110 is unbypassed, providing some degenerative feedback. Output is developed across a parallel-tuned resonant circuit composed of the primary of T101 and C109. R114 and C107 form a decoupling network, while the latter also places the upper end of the primary of T101 at AC ground.

The secondary of T101 presents a signal at low impedance which is coupled through noise blanker gate Q104 to the primary of T102, where it is stepped back up. C112 resonates the secondary of T102. In the transmit condition, the noise blanker gate is held in the on condition at all times.

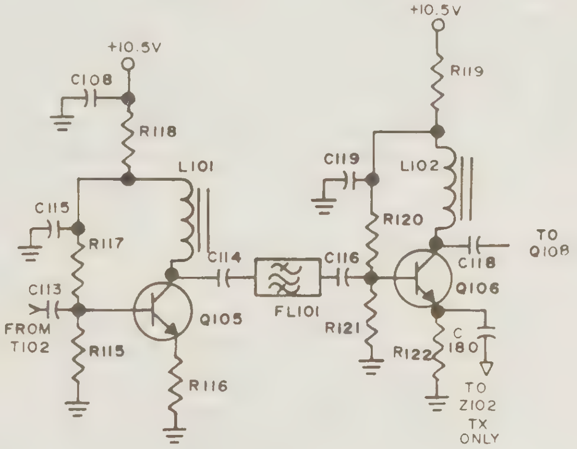


FIRST I.F. AMPLIFIER & NOISE BLANKING GATE

FILTER BUFFER AND FILTER

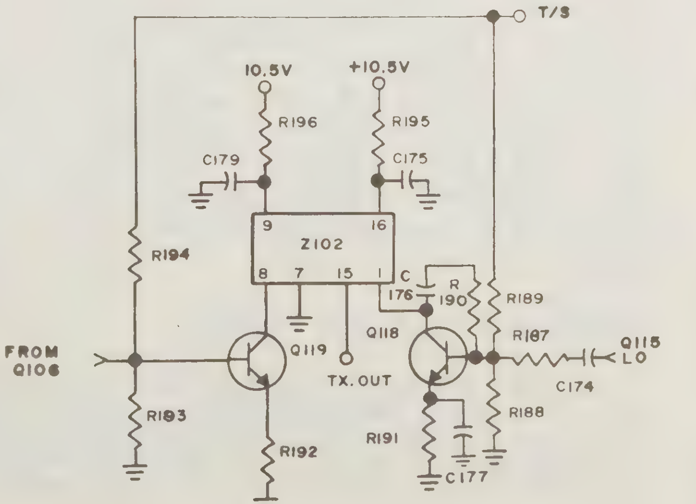
From T102, the signal is coupled to the filter buffer amplifier, Q105, via blocking capacitor C113. Bias is developed by R115 and R117, while emitter resistor R116 is unbypassed. Output is taken across collector load L101 and coupled through blocking capacitor C114 to the first IF filter, FL101. R118, C108 and C115 are a decoupling network.

From filter FL101, the signal is coupled through C116 to the base of post filter amplifier Q106, which is biased by voltage divider R120 and R-121. In the transmit condition Q106 acts as an emitter follower, with its output coupled via C180 and R197 to the transmit mixer driver, Q119.



TRANSMIT MIXER

As mentioned above, the signal from the post filter amplifier, Q106, which is now a single sideband, suppressed carrier signal at the IF frequency of 9 MHz, is coupled to transmit mixer driver Q119. This stage has conventional bias, an unbypassed emitter, and its output taken from the collector through a winding of Z102. This winding is decoupled by R196 and C179. Bias supply comes from the T/S line from the control circuitry,



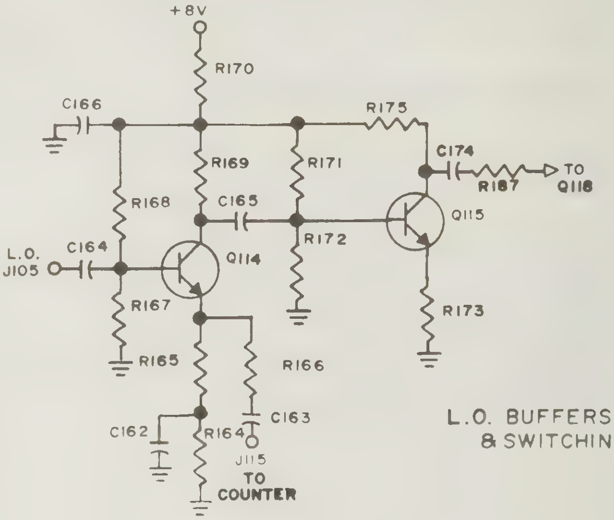
TX MIXER, DRIVERS

which will be covered later.

Local oscillator energy is coupled to the base of the other driver, Q-118. The gain of this stage is determined by the ratio of feedback resistor R190 to input resistor R187; C174 and C176 are merely DC blocking capacitors. The emitter of the stage is bypassed conventionally. Output is taken from the collector through a winding of the transmit mixer, Z102, a doubly balanced mixer assembly identical to that used for A101.

L.O. drive is connected to Z102 through an amplifier/buffer network contained on the exciter board also. Signal from the local oscillator board is applied to buffer Q114 via C164. Output is taken from the R165 portion of the emitter resistance, and coupled through R166 and C163 to the counter board, to provide the signal for the front panel digital dial, while the remainder of the emitter resistance, R164, is bypassed by C162.

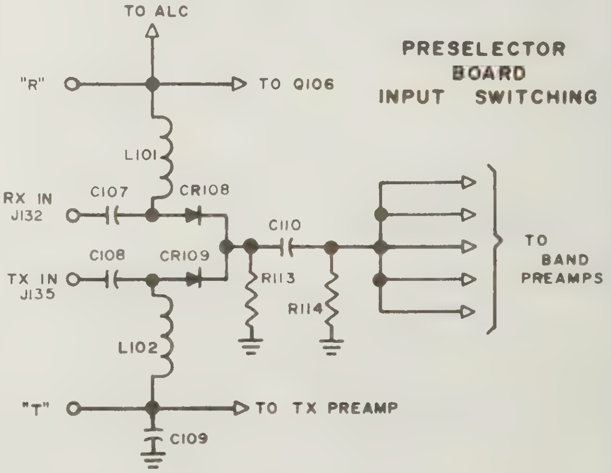
Signal is also taken from the collector of Q114, coupled through C165, and applied to the base of Q115. The emitter resistor of this stage is unbypassed, creating some degree of degenerative feedback, and the output is taken via C174 and R187 to the mixer driver, Q118.



L.O. BUFFERS & SWITCHING

PRESELECTOR--INPUT SWITCHING

Output from transmit mixer Z102 which is now at the desired operating frequency, is coupled to the same preselector board used in the receive portion of the transceiver. It is coupled via C108 and C110 to the input of the band preamplifier filters when the "T" line is high, biasing switch CR109 on through L102 and R113. Since the "R" line is low under transmit conditions, the voltage dropped across R113 reverse biases receive switch CR108, disconnecting the receiver input from the preselector.

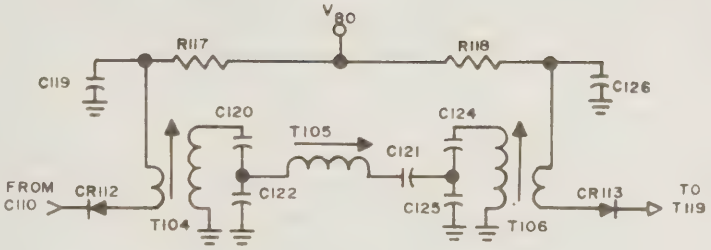


PRESELECTOR BOARD INPUT SWITCHING

PRESELECTOR--BANDPASS FILTERS

Since operation of all of the individual band filters is identical, only one, the 80 meter version, is discussed here. These filters are selected by the BAND switch on the front panel of the transceiver, utilizing the control lines from S103 on the Bandswitch/Low Pass Filter board.

The voltage from the 80 meter control line is coupled through isolating resistor R-117, the primary of T104, and CR112 to the output of the switching circuitry discussed in the previous section, where it is provided with a ground return path through R-114. C119 provides a ground for the signal at one end of the primary of T104. The signal is then passed through diode switch CR112 to T104, while the voltage drop across R114 reverse biases the corresponding diode switches in the other filters.

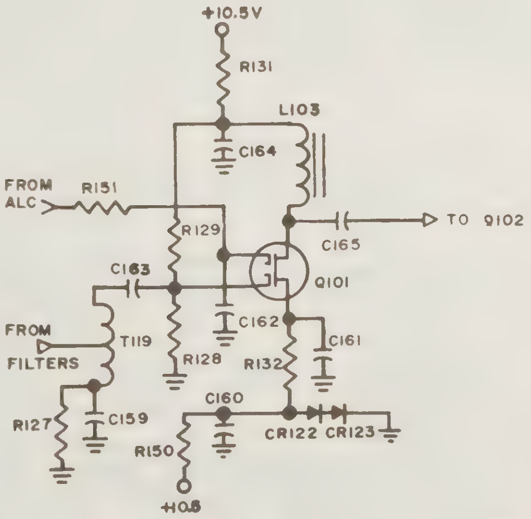


80 M BAND FILTER

The secondary of T104 is resonated by the C120-C122 series combination, and they also act as an impedance transformation voltage divider. T105 and C121 form a series resonant circuit, and its output is coupled to C124-C125, which again transforms the impedance of the signal line, and also resonates with output transformer T106. The combination of the two parallel resonant circuits and one series resonant circuit forms a bandpass filter of excellent characteristics. Output switching is provided by R118, in conjunction with CR113, in a manner identical to the input switching scheme.

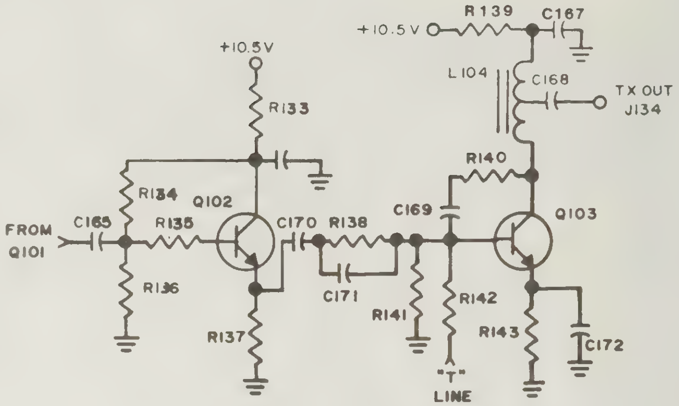
The output of the preselector filters is coupled to the preamplifier stage, Q101, via autotransformer T119 and coupling capacitor C163. R127 provides a DC path to ground for the switching diode, while C159 effectively grounds the bottom of T119 for the signal.

Bias for Q101 is provided by both the source circuitry and the R128-R129 combination. So far as gate two is concerned, the bias is determined by the source network and what ALC is applied to it from the ALC circuitry, to be discussed later. CR122 and CR123, biased in the forward direction by the voltage applied to R150, provide a fixed drop of approximately 1.4 volts, while additional bias is supplied by the drop across source resistor R132. C160 and C161 are both bypasses.



Output, via C165, is taken from Q101 and applied to transmit preamplifier Q102, which is a conventionally biased emitter follower stage. This in turn is coupled via C170 to frequency-equalizing network R138-C171, and then to the base of feedback amplifier Q103. Gain of this stage is determined by the ratio of R140 to the impedance of the R138-C171 combination. Bias is applied only in the transmit condition, from the "T" line, via voltage divider R141-R142. The emitter circuit is composed of R143 bypassed by C172.

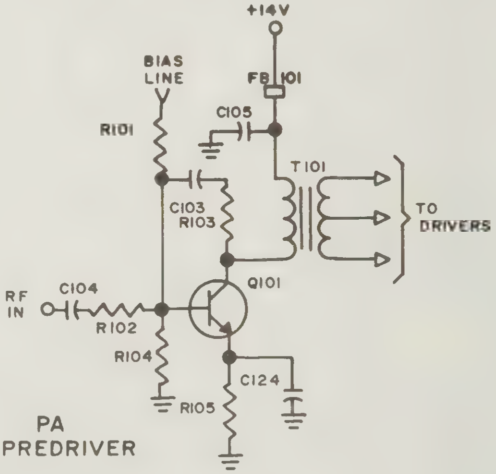
Output is taken from a tap on autotransformer L104 and coupled through C168, providing a low impedance signal that can be coupled through a coaxial cable to the power amplifier board with minimum loss. C167 and R139 form a supply line decoupling network.



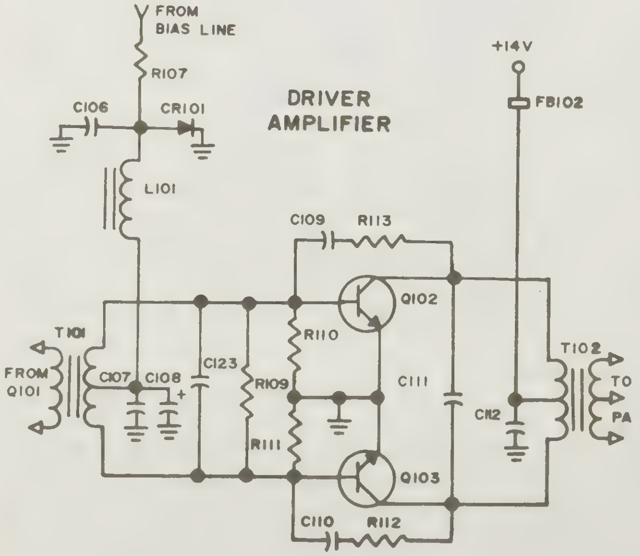
TRANSMIT PREAMP

POWER AMPLIFIER

From the preselector, the transmit signal is applied to PA predriver Q101 on the power amplifier board via coupling capacitor C104. Since Q101 is a feedback stage, its gain is set by the ratio of feedback resistor R103 to input resistor R102, while bias is taken from the PA bias line via voltage divider R101-R104. The emitter circuit is conventional, and the output is taken via T101, which provides a push-pull signal for the following driver stages. C105 and ferrite bead FB101 provide decoupling for the collector supply line.



PA
PREDRIVER



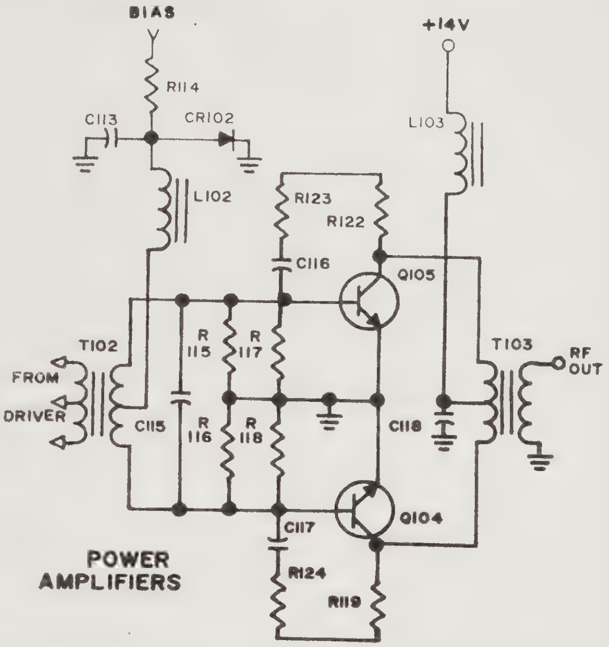
DRIVER
AMPLIFIER

The driver amplifiers, fed by transformer T101, are supplied bias via the center tap of that transformer and RF choke L101. R107 and CR101 form a forward-biased diode regulator circuit to control the amount of bias supplied. C106, C107 and C108 are bypasses.

The C123, R109, R110 and R111 network assures the correct load impedance on the secondary of T101. The R113-C109 and R112-C110 networks provide feedback for the stages. C111 helps to provide proper frequency response at the primary of the interstage transformer, T102, while C112 and FB102 provide required decoupling of the supply line.

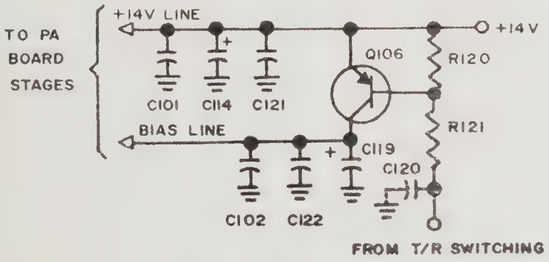
Power amplifier stages Q104 and Q105 both are also driven in push-pull, from the secondary of transformer T102. Bias is supplied in a manner similar to that of the driver stages, and again, a network composed of C115, R115, R116, R117 and R118 assure proper termination impedance for the secondary of T102. Feedback networks are provided for these stages also, and output appears across the secondary of T103. C118 and RFC L103 provide supply decoupling.

The bias line for the PA board is controlled by Q106. Normally, the base of this device is held at the +14 volt supply potential through R120, thus cutting it off. When, however, the transmit mode is activated, the T/R switching circuits ground the lower end of R121, thus lowering the voltage on the base of Q106 and allowing it to saturate. This energizes the bias line and all circuits supplied from it.



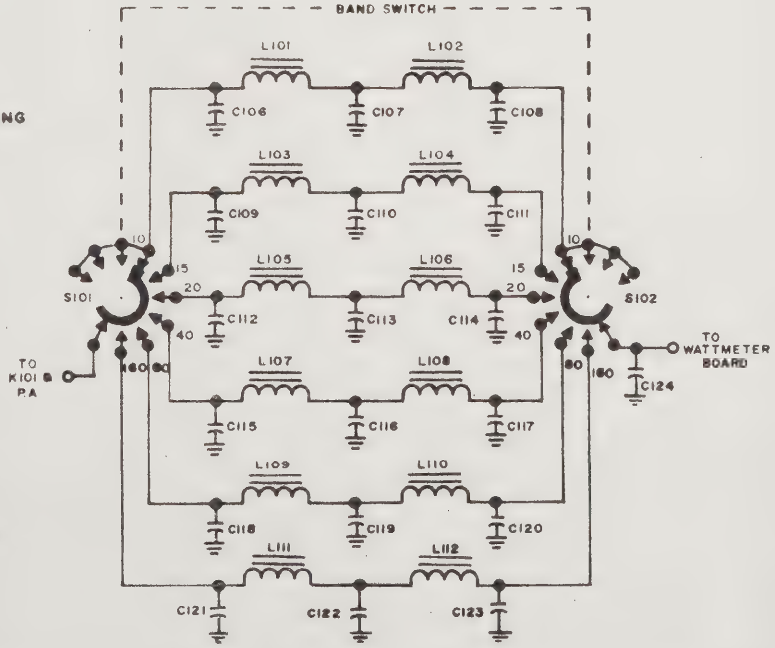
POWER AMPLIFIERS

Capacitors C102, C122 and C119 on the bias line, and C101, C114 and C121 on the +14 volt supply line assure good bypassing at all frequencies from audio to radio.



PA BOARD BIAS

The output from the power amplifier is coupled to switch S101 on the bandswitch/low pass filter board, and thence through the appropriate low pass filter for the band selected, and out through S102 to the wattmeter/SWR board (covered later). Each of the low pass filters on this board is a five pole type, designed to greatly attenuate out-of-band harmonics and spurious outputs.



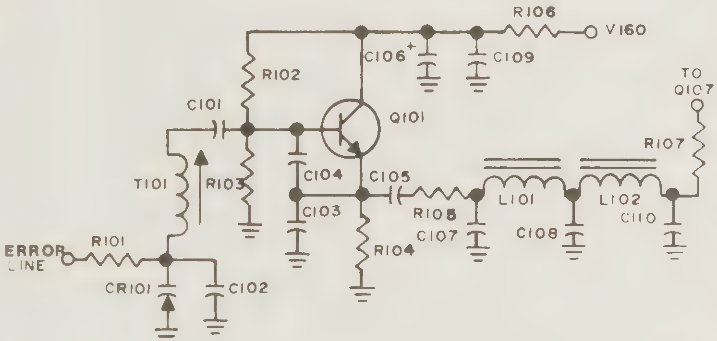
Oscillator Theory

Band Voltage Controlled Oscillators.......5-1
Permeability Tuned Oscillators (PTOs).....5-3
PTO Switching.............................5-4
Local Oscillator Mixer....................5-5
Carrier Oscillator........................5-7
CW Sidetone Oscillator....................5-7
-5 Volt Oscillator/Regulator..............5-8

BAND VOLTAGE CONTROLLED OSCILLATORS

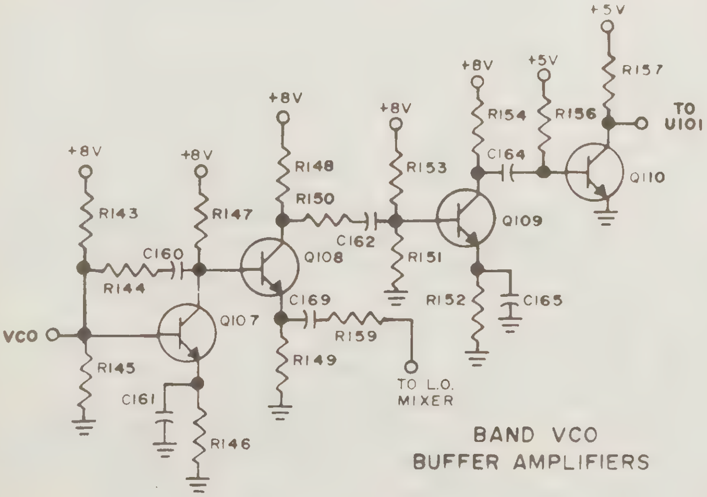
Each of the six bands has its own voltage controlled oscillator (VCO) that operates at a frequency 14.5 MHz higher than the low edge of the band itself. Each of these VCOs is the same as the other, with the exception of the values of the frequency sensitive components, so only one will be discussed here, that for the 160 meter band.

The oscillator circuit is derived from the basic Colpitts oscillator, with C103 and C104 providing the tapped capacitance for feedback. T101 and the parallel combination of varicap CR101 and C102 provide the tuned circuit. Voltage applied by the error line through isolation resistor R101 tunes the varicap. C101 is a blocking capacitor, and bias is developed by voltage divider R102-R103. Output is taken through blocking capacitor C105 and series resistor R105, and is filtered by a low pass filter composed of inductors L101 and L102, with C107, C108, and C110.



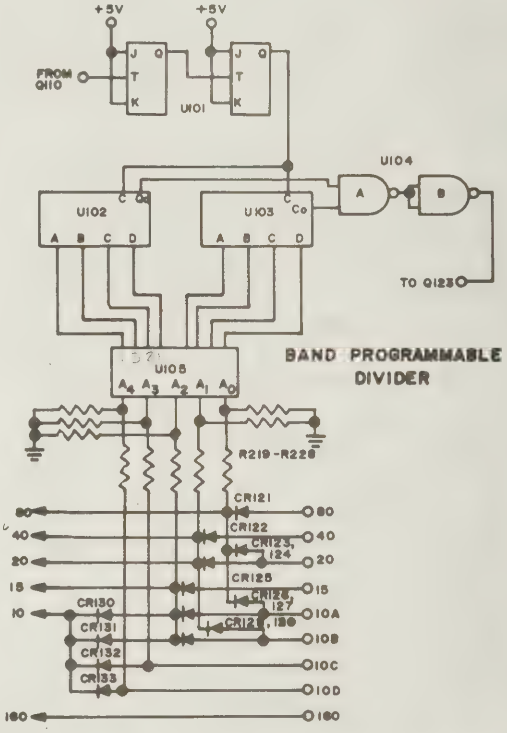
BAND VCO

The output of the filter is coupled through series resistor R107 to a line common with the other five oscillators, and to the base of summing amplifier Q107. R144 and C160 provide a feedback path from collector to base of this stage, so the gain of Q107 is determined by the ratio of R144 to R107. (See the "Servicing Hints" section of this manual for a discussion of feedback amplifiers). Bias for Q107 is conventional, as is the emitter circuitry. The collector of Q107 is directly coupled to the base of Q108, making them into a high-gain compound pair. Output to the L.O. mixer is taken from the emitter of Q108 through C169 and R159, while an additional output is taken from the collector through R150 and C162, and applied to the base of amplifier Q109. This stage then drives level translator Q110 through C164, and bias is applied through dropping resistor R156. The output of Q110, now a square wave at a TTL-compatable level, is directly coupled to the prescaler of the band phase locked loop. (See the "Servicing Hints" section for a discussion of the PLL.)



BAND VCO BUFFER AMPLIFIERS

U101, a dual J-K flip flop, is used as a divide-by-four prescaler for the divider chain. Its output is coupled to the clock inputs of U102 and U103 in parallel. The outputs of U102 and U103 are gated through U104A to produce an output pulse when the counter has completed its counting cycle; this pulse is then inverted through U104B, and used to drive the phase detector driver, Q123.

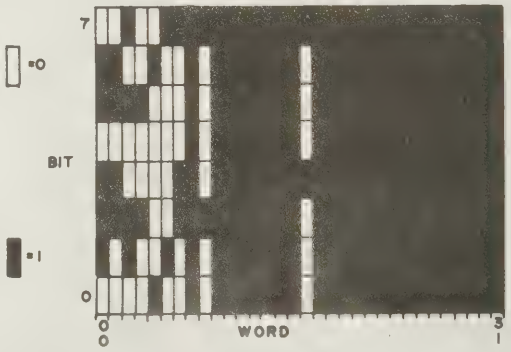


BAND PROGRAMMABLE DIVIDER

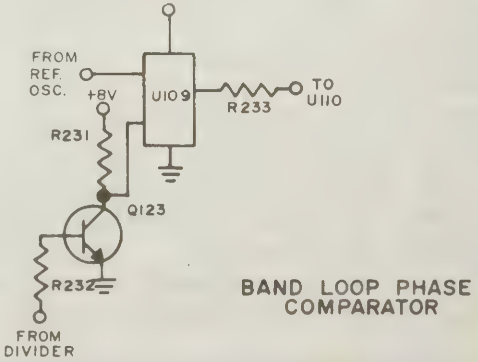
Programming of counters U102 and U103 is accomplished by the diode matrix composed of CR121 through CR129, and memory U105. For 160 meter operation, all five input lines to U105 are low, or binary 00000; for 80 meters, the coding is 00001; for 40, 00010; for 20, 00011; for 15, 00100; for 10A, 00101; for 10B, 00110; for 10C, 01000; and for 10D, 10000. Band program memory U105 decodes these addresses into the bit combinations necessary to set the required division ratios into counters U102 and U103, as shown in the accompanying memory map of U105.

Additional diodes (CR130 through CR133) are connected to the 10 meter control lines which provide a common 10 meter bus, regardless of which segment of the band is selected by the BAND switch, and is used to energize the 10 meter VCO and 10 meter L.O. filter.

U105 MEMORY MAP



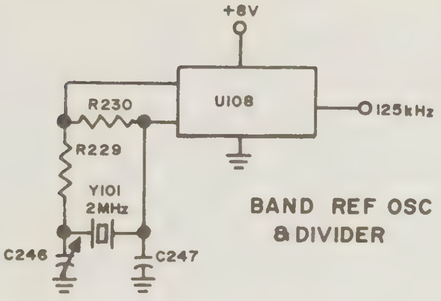
The output of the programmable divider is coupled through R232 to the base of driver Q123. Its output is coupled to the phase comparator, U109. The other input to the phase comparator is the 125 kHz signal from U101B, the loop reference oscillator and divider. Y101 controls the frequency of the oscillator at 2 MHz, with C246 providing a means of adjusting the calibration.



BAND LOOP PHASE COMPARATOR

Output of the phase comparator, U109, is coupled through R233 to the inputs of two amplifier stages, U110A and U110B. Capacitor and resistor combinations C248-R233 and R239-C249 form the loop filter. For 160 through 15 meter operation, U110A is used as the error amplifier, with its output driving the varicaps in those band oscillators; for the 10 meter band, U101B, with higher gain, is used to drive the varicap of that band's oscillator. More gain is required because the higher divider numbers also reduce the loop gain.

In this manner, the band VCOs are constrained to oscillate on precise multiples of the crystal controlled reference frequency of 125 kHz. The exact multiple is the total division performed on the VCO frequency, including that of the prescaler.

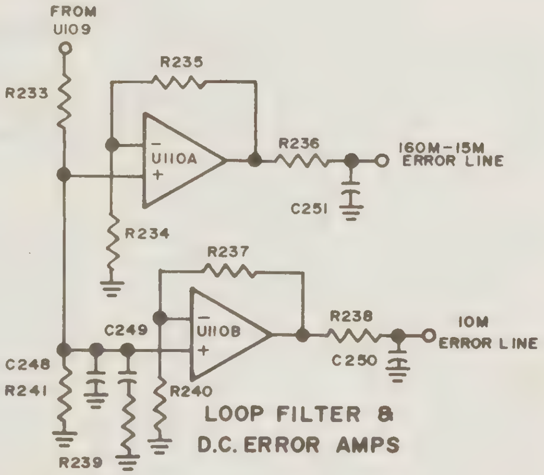


BAND REF OSC & DIVIDER

PERMEABILITY TUNED OSCILLATORS

There are two identical permeability tuned oscillators in the ASTRO 102BX. Only one is described here.

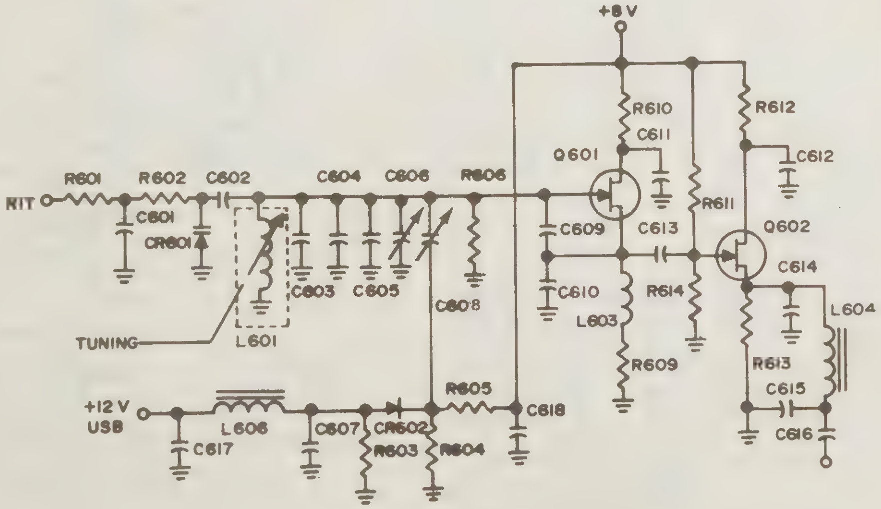
The permeability tuned oscillator is the primary means of varying the frequency of the transceiver. It is a Colpitts circuit, and the tuned frequency is determined by L601 in parallel with the appropriate capacitors, CR601, C603, C604, C605 and C606. The frequency tuned is 5.0 to 5.5 MHz, and this combination of capacitors is chosen to provide temperature compensation for low drift, and must be replaced with exactly the same types to maintain this stability, if replacement becomes necessary.



LOOP FILTER & D.C. ERROR AMPS

R606 provides a DC return for the gate bias of Q601, which is developed by R609. Feedback to sustain oscillation is achieved by the C609-C610 divider combination. The drain is decoupled to ground by R610 and C611, while the output is taken from the source to drive FET buffer Q602.

The bias generated by source resistor R613 of Q602, whose value is set by the desired output impedance, is greater than is desirable for this FET; hence, a positive voltage to offset this is developed



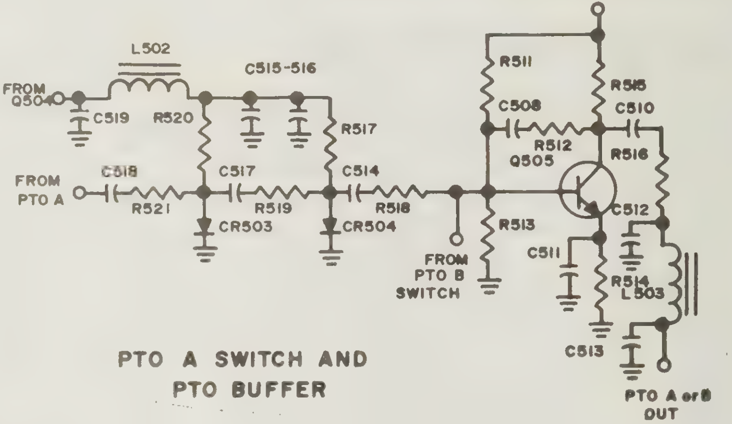
PERMIABILITY TUNED OSCILLATOR

by the R611-R614 combination. Output is taken via low pass filter C614-L604-C615 to assure good spectral purity.

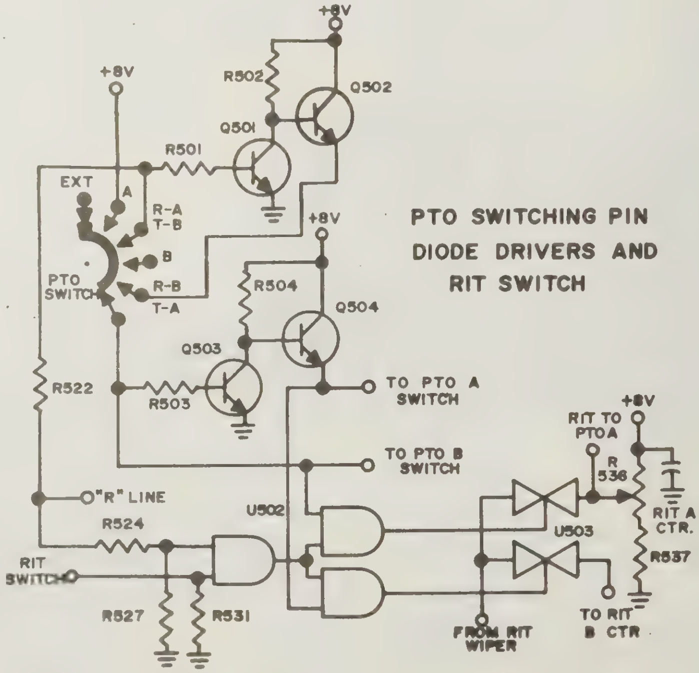
When operating in the USB mode, the required frequency shift is set by C608, which is switched in parallel with the tuned circuit by CR602.

When RIT is used, the required shift is obtained by varactor diode CR-601, connected in parallel with the tuned circuit by blocking capacitor C602. The DC tuning voltage for this varactor is applied through the R601-R602 combination, and is decoupled by C601. C606 is a calibration control.

PTO SWITCHING



PTO A SWITCH AND PTO BUFFER



PTO SWITCHING PIN DIODE DRIVERS AND RIT SWITCH

The switches are used for the PTO outputs; only one will be described here, as both are identical and the description of one will apply equally to both.

A control voltage, the generation of which will be discussed later, is applied to decoupling RF choke L502 from Q504. C519, C515 and C516 are bypass capacitors. This voltage is applied to PIN diode attenuators CR503 and CR504 through isolating resistors R520 and R517, causing the diodes to present a low resistance to ground. Since the signal from the PTO passes through R521 and R519 (as well as blocking capacitors C518 and C517, it is dropped by these two resistors, with essentially none passing through to the base of the buffer, Q505. Thus, when this control voltage is removed, the PTO output is allowed to pass to the buffer. Q505 is a summing amplifier, with its gain determined by the ratio of R512 to R518 (and others, when the other section is in use). Bias and emitter circuitry are conventional, and output is taken from the collector through low pass filter C512-L503-C513 to assure that no harmonics or spurs are passed on to the mixer. In the off condition, the switch provides greater than 100 dB of attenuation to the unwanted PTO output.

The control signal for the PTO switching circuitry is derived from the front panel PTO switch. In the PTO A position, +8 volts is applied to the PTO B switching circuitry directly through the switch contacts, thus disabling PTO B. It is also applied to the base of Q503, causing it to saturate; since Q504 is direct coupled to its collector, the latter is then turned off and no voltage appears on the PTO A control line, allowing its signal to pass and appear at the output.

In the PTO B position, no voltage is applied to the PTO B line, allowing it to function, and Q504 is turned on, allowing voltage to appear on the PTO A control line, disabling PTO A.

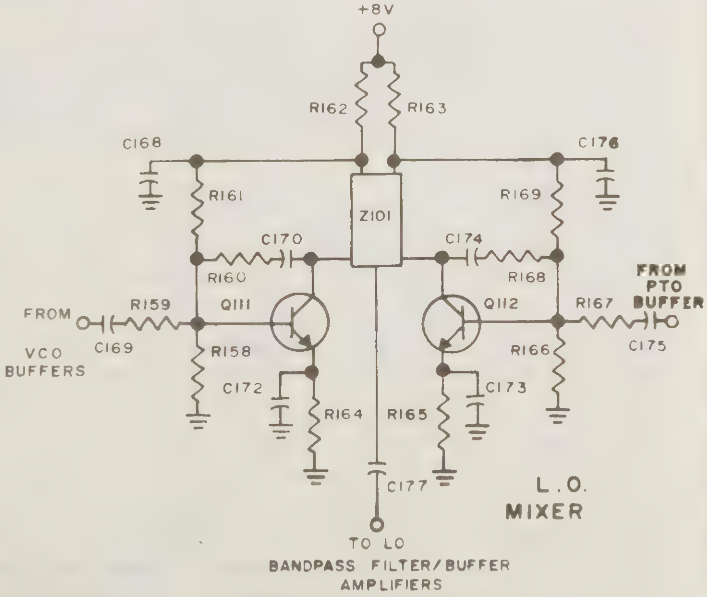
For split operation, "R" line voltage is used (through the PTO switch on the front panel) to drive the PTO B control line and the Q503-Q504 combination. Thus, which PTO is turned on depends upon whether the "R" line is high or low (receiver on or off). For reverse split operation Q501 and Q502 provide an inverted "R" line voltage (+ for transmit, 0 for receive) which is used instead.

When the RIT switch on the front panel is on, and the transceiver is in the receive condition ("R" line high), AND gate U502A is on, and its output is high. This high is coupled to one input of U502B and U502C. The other inputs of each of these AND gates is connected to the PTO A and PTO B control lines, respectively. Thus, when the PTO B control line is high (and the PTO A line is low), analog switch U503A is turned on and the voltage from the RIT control wiper is applied to the RIT input of PTO A. R536 and R537 provide center adjustment for the RIT circuitry in the PTO.

When the PTO A control line is high, the remaining section of the AND gate is on, turning on U503B, and supplying the RIT control voltage to PTO B. A center adjustment similar to that for the A PTO is included in the circuit although not shown in the diagram, for simplicity.

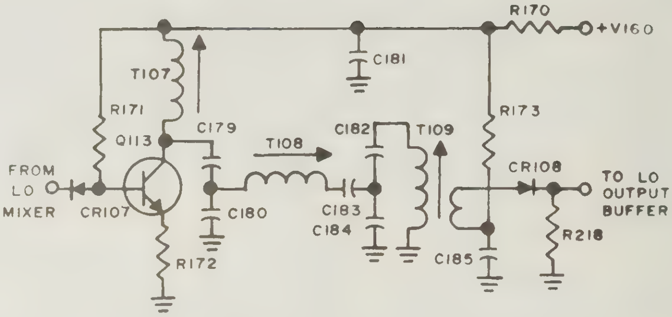
LOCAL OSCILLATOR MIXER

Output from the band synthesizer (VCO buffer) is applied to mixer driver Q111 through blocking capacitor C169. Q111 is a feedback amplifier whose gain depends on the ratio of R160 to R159, and its output is applied to one input of the L.O. mixer Z101, which is a doubly-balanced mixer identical to others used in this transceiver. Also, the output of the selected PTO is applied to Q112 through blocking capacitor C175. This stage too, is a feedback amplifier, with its gain determined by R167 and R168. Its output is similarly coupled to Z101.



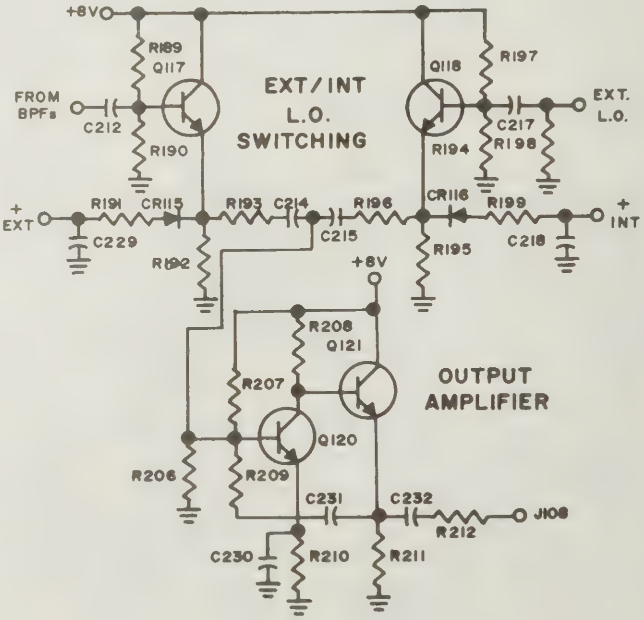
Output of Z101 is at the operating frequency plus 9 MHz, and is processed through one of six bandpass filter-amplifier combinations. Since all six are identical in operation, only the 160 meter version will be discussed here.

Collector and switching power comes from the 160 meter control line. R171 supplies bias for amplifier Q113, and also biases switch CR107 on; the voltage drop across its ground return resistor (not shown in the simplified diagram) reverse biases similar diode switches in the other bandpass filter circuits.



LO BANDPASS FILTER/BUFFER

Q113 operates without an emitter bypass capacitor, and is therefore a bit degenerative; its collector is composed of a parallel resonant circuit formed by T107 and the C179-C180 combination. Output is taken from the low impedance tap between these two capacitors, and coupled through series resonant circuit T108-C183, then stepped back up to higher impedance by the output parallel resonant circuit T109-C182-C184. Output switching is via CR108, biased on through R173, the secondary of T109, and R218. C185 is a bypass for the signal. Again, the voltage drop across R218 reverse biases the output switching diodes in the other filters.

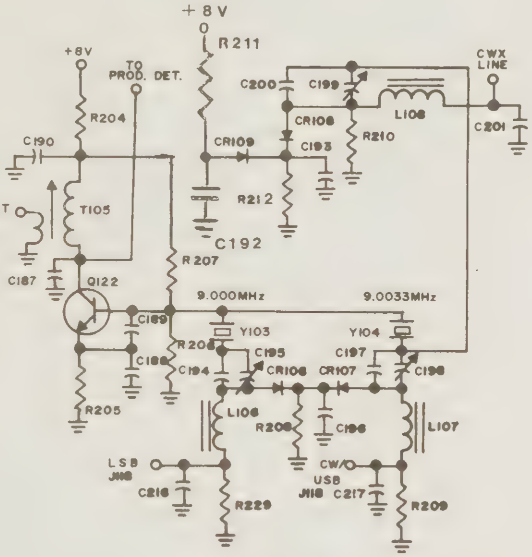


When the PTO switch on the front panel is in the EXTernal position, a voltage is applied through R191 to CR115, and thence to the emitter of L.O. buffer Q117. This reverse biases this stage's base—emitter junction, preventing internally-generated L.O. energy from being passed on. At the same time the voltage drop across R195 reverse biases CR116 (since the INT line is low) and Q118 acts as a buffer for externally supplied L.O. signal, and it is passed on through R196 and C215 to the L.O. output amplifier, Q120.

When the PTO switch is in any other than the EXT position, the INT and EXT voltages are reversed, and internally generated L.O. signal is passed on to Q120.

Q120 and Q121 are directly coupled, and amplify the L.O. signal selected by the switching circuit. The emitter of Q120 is wired conventionally, while R206 and R207 are the bias divider. Feedback from the emitter of Q121 to the base of Q120 is provided by the R209-C231 network. Output is taken from the emitter of Q121 through C232 and R212, and routed to the exciter board.

CARRIER OSCILLATOR



CARRIER OSCILLATOR

Since the ASTRO 102BX utilizes only one crystal filter, the carrier oscillator frequency must be shifted appropriately when selecting USB, LSB or CW modes, so that the signal will fall within the pass band of the filter. Only one oscillator is used, with a crystal switching scheme used to select the proper crystal, depending upon the mode selected.

Q122 is the oscillator, with bias provided by R206 and R207. Feedback is provided by C188 and C-189. R205 is the emitter resistor. The collector load is comprised of T105 and C187, which form a parallel resonant circuit. R204 and C-190 form a decoupling network.

In the LSB mode, a + voltage is applied through L106 which forward biases diode CR106, with R208 serving as a ground return. In this manner, C196 effectively grounds the lower end of C194 and C195, placing Y103 into the circuit between Q122's base and ground; the oscillator then operates at Y103's frequency, as trimmed by C195.

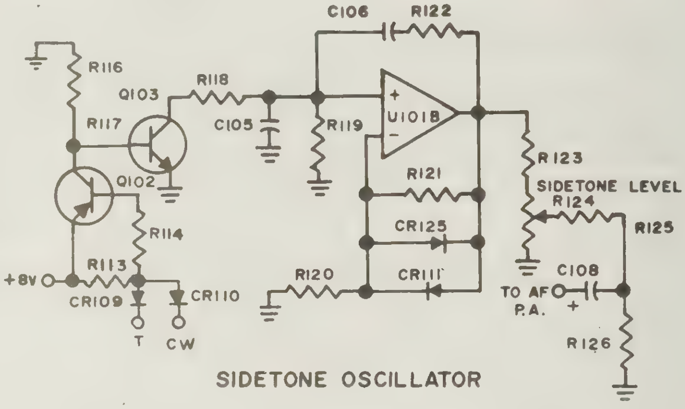
At the same time, the voltage drop across R208 reverse biases CR107, preventing Y104 from having any effect on the oscillator frequency. In the USB or CW receive modes, the + voltage is applied via L107 to CR107, forward biasing it in the same manner as CR106 in the LSB mode. CR106 is now reverse biased, and Y104 controls the frequency of oscillation. The exact frequency is set by C198.

When transmitting in the CW mode, an offset is desired to place the signal into the passband of the IF filter. In the ASTRO 102BX, this offset is approximately 800 Hz from the receive condition, and is provided by "pulling" the frequency of Y104. In transmit, the CWX line is high, forward biasing CR108 and reverse biasing CR109. C199 and C200 now affect the frequency of Y104, and C199 is adjusted during alignment to provide the desired frequency offset.

CW SIDETONE OSCILLATOR

When operating in the CW mode, a sidetone is provided in the transceiver's loudspeaker so that the operator may monitor his own sending. This signal is generated by CW sidetone oscillator U101B. The frequency of the sidetone is determined by the positive feedback path for U101B, that is, C-106, C105, and R118. The gain of the stage is determined by the inverting feedback path, R121, CR125, CR111 and R120, with the diodes providing a leveling action for the oscillator output. The output is applied to sidetone level control R124, and thence to the audio frequency power amplifier via coupling capacitor C108.

Keying of the sidetone oscillator is accomplished by Q102 and Q103. When in the CW mode, CR110 is reverse biased, but the base of Q102 is held low by the "T" line, maintaining Q102 in a conducting state. When the key is closed, the "T" line goes high, allowing the base of Qlo2 to rise, and cutting off conduction in R116 in the collector circuit

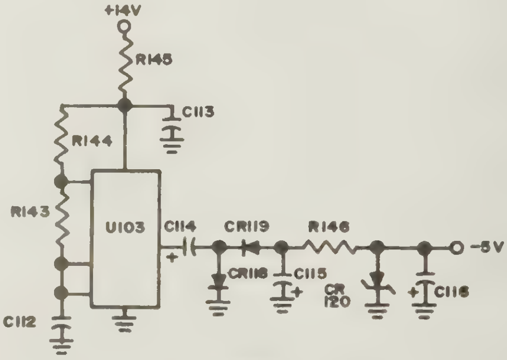


SIDETONE OSCILLATOR

Thus, when the key is closed, and the MODE switch is in the CW position the base of Q103 is low due to the fact that Q102 is not conducting, and it does not conduct. This allows U101B to oscillate. When the key is opened, the "T" line goes low, CR109 conducts, and the base of Q102 is grounded. It thus conducts, creating a voltage drop across R116, and biases Q103 on. This prevents U101B from oscillating.

-5 VOLT OSCILLATOR/REGULATOR

The negative five volts DC required by various portions of the transceiver circuitry is provided by oscillator U-103, a timer IC connected in the astable mode. The frequency of oscillation is set by the R143, R144 and C112 combination. Output is taken through capacitor C114, and applied to a voltage doubler composed of CR118 and CR119. C114 and C115 are an integral part of the doubler circuit, charging alternately on different half-cycles of the oscillator output. Their individual voltages add during discharge, providing the required voltage doubler action.



—5V OSCILLATOR AND REGULATOR

The negative voltage thus generated is regulated by the combination of dropping resistor R146 and zener diode CR120, in a conventional zener regulator circuit. Output is filtered by C116.

Ancillary Circuits

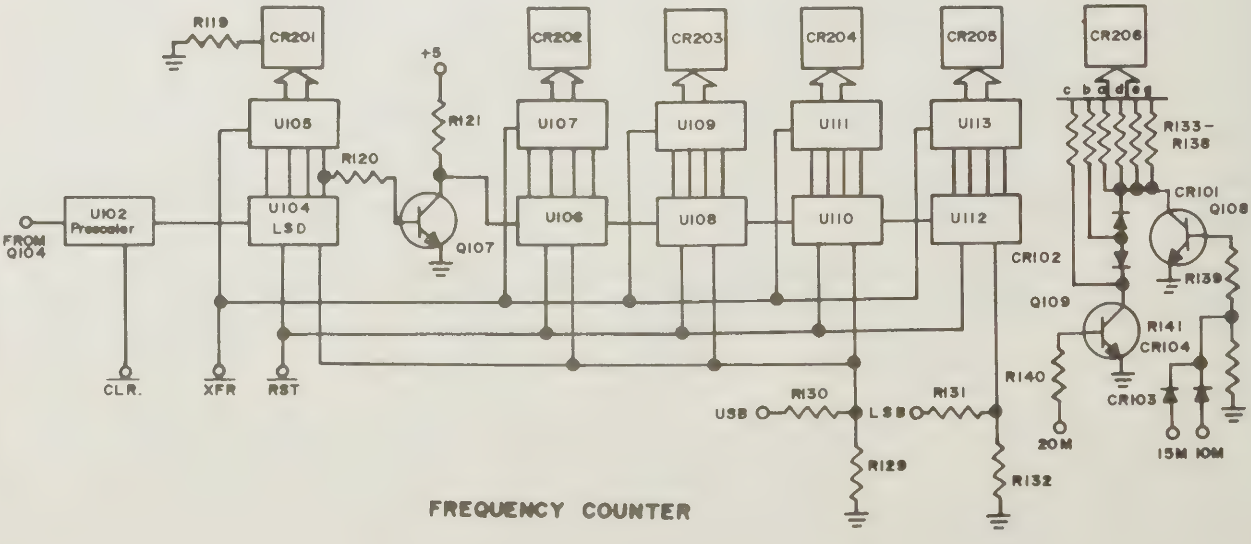
Frequency Counter/Digital Dial......................6-1
10.5 Volt Regulator...............................6-4
Wattmeter/SWR Detector............................6-4
Automatic Level Control (ALC).....................6-5
VOX...6-6
Keying Control....................................6-7
CW Timer..6-8
Noise Blanker.....................................6-9
Passband Tuning Indicator.........................6-10

FREQUENCY COUNTER

The ASTRO 102BX uses an accurate, crystal controlled time base and a frequency counter to provide the operator with a digital display of the transceiver's operating frequency. This counter measures the local oscillator frequency, corrects for the fact that this is 9 MHz above that which the transceiver is receiving or transmitting, and displays the results.

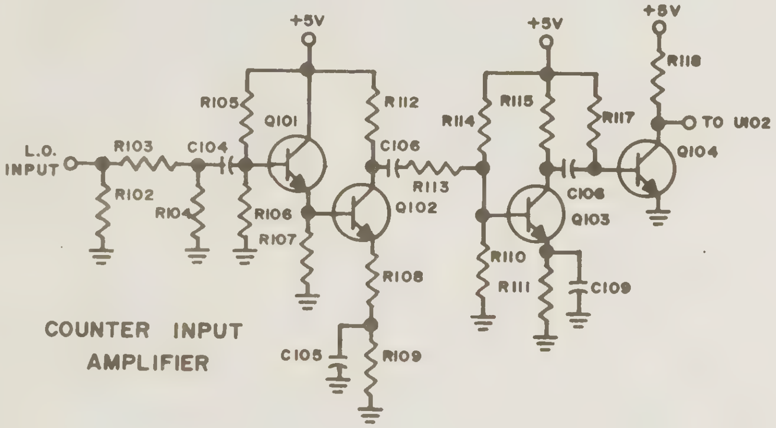
The incoming local oscillator signal, taken from the emitter of Q114 on the exciter board, is amplified and processed by the counter input amplifier. R102, R103 and R104 form an attenuator network, used to reduce any spurious signals generated in the counter, preventing them from being coupled back into the receiver circuitry. The overall design of the counter includes extensive planning to prevent "counter birdies"--an unfortunate phenomenon in all digital dial radios. This includes the solid shielding box and cover that the counter is enclosed in, feedthrough capacitors on leads penetrating this shield, high isolation amplifiers, input attenuator, etc. The results are that the unavoidable spurs are reduced to insignificant levels. The technician is cautioned to exercise particular care in maintaining the integrety of this shielding.

C104 couples the signal to the base of the first of two devices connected as a compound pair, Q101 and Q102. Bias for Q101 is provided by conventional voltage divider R105-R106, and the drop across its emitter resistor, R107, provides bias for Q102. The emitter resistance of Q102 is divided into two parts, R108 and R109, with the latter bypassed by C105. This arrangement compensates for the inherent 6 dB-per-octave rolloff in transistor gain, thus levelling its response over the range desired. Output from Q102 is taken via the collector, coupled through C106 and R113, and applied to the base of Q103. This stage is conventional in all respects, with R114



FREQUENCY COUNTER

and R110 forming the bias network and R111, bypassed by C109, the emitter circuit. The output of Q103, now gain-levelled and amplified, is applied to the base of Q104 through C108. Since the desired output of this stage is a square wave suitable for driving the counter prescaler, and linearity is no longer of importance, R117 suffices to provide bias for Q104. The collector is directly coupled to the input of the prescaler.



COUNTER INPUT AMPLIFIER

U102, a divide-by-ten counter, reduces the frequency of the incoming signal from the input range of approximately 10.8 to 39 MHz (9 MHz above the transceiver operating frequency) to 1.08 to 3.9 MHz for the input of the counter proper. It is cleared at the end of each count cycle by the negative-going pulse on the clr line. This line provides the necessary gate function to drive the counter proper.

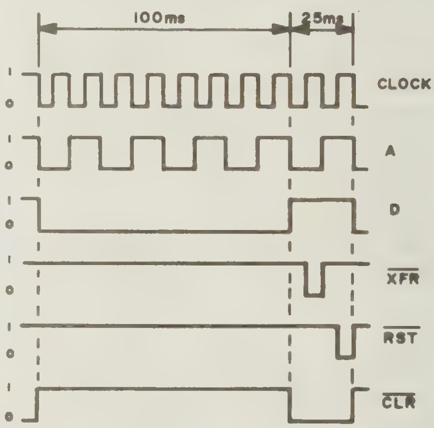
The counter chain is conventional, with two exceptions; the output of the prescaler is still too high for reliable operation of the CMOS devices, particularly complex devices such as counters, on a 5 volt supply; hence, in this decade, a low-power Schottky device is used. This counter, the least significant figure decade (100 Hz on the display), counts using the conventional BCD code, and its four output lines drive U105, a combined storage latch, decoder, and seven-segment display driver. When the end of the count cycle is reached, the xfr line is brought momentarily low causing the latches in U105 to accept the data being fed to it by U104, and to store it until updated at the end of the next counting cycle. This data is then decoded, and the proper segments of the display, CR201, are activated, displaying the count. CR201 is a common anode type of display, so grounding the input pins for the appropriate segments will cause them to light. Current is limited to the correct value (15 mA) by internal circuitry of U105. R119 is used to provide a path to ground for the decimal point of CR201, and to limit its current, so that it is lighted at all times the radio is on.

The second somewhat unusual feature of the counter is that the output of the LSD decade is not taken from the counter "carry" output, for the pulse presented here is much too narrow for the following, CMOS decade. Instead, output is taken from the Q_D terminal, through current limiting resistor R120, and applied to the base of Q107. The collector, with pull up resistor R121, is connected to the input of the next decade counter, U106. The pulse thus obtained is much wider than that presented at the "carry" output, and is inverted to the correct polarity.

The subsequent decades of the counter are all of the CMOS family--U106, U108, U110 and U112. Each has a "carry" output which is used to drive directly the "up count" input of the subsequent decade. Each has four BCD coded data output lines which are connected to the latch/decoder/driver IC corresponding to that decade, i.e., U107, U109, U111 and U113. Each of these in turn operates the appropriate display indicators. The latches are all strobed once per count cycle to update the display, in the same manner as described for U105.

In order to read the incoming frequency correctly, to the nearest 100 Hz, the counter would have to be gated "on" for 10 milliseconds; however, since we use a prescaler with a divide-by-ten ratio, we must extend the count cycle time to 100ms. This is accomplished by the clr line of U102, the prescaler. It operates, and passes on its output pulses, only when that line is high; thus, for the time it is low, the counters stop, but store their present count while the latches are strobed, thus passing that count to them (and the displays), and then are preset by the rst line, ready to start the next count cycle.

COUNTER TIMING CHART



Unfortunately, it takes a finite time to (a) stop the counting, (b) transfer the count to the latches, (c) clear the counters and (d) start the next count cycle. This is accomplished, as shown in the accompanying counter timing chart, during the 25 ms immediately following the counting portion of the counting cycle. This overall cycle translates to a rate of eight updates of the display per second.

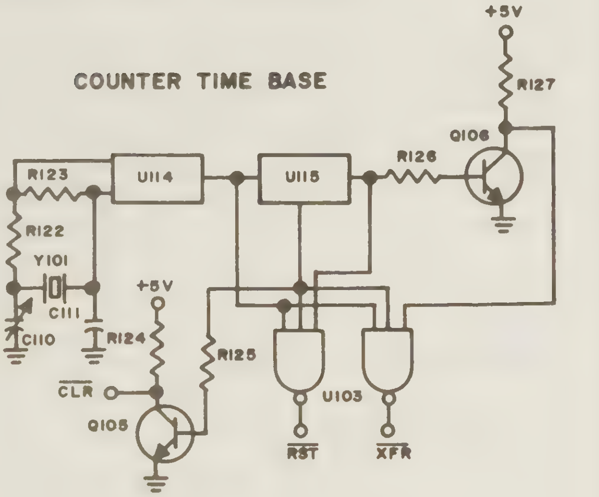
For a laboratory-style frequency counter each decade would be reset to zero (BCD 0000) at the end of each complete count cycle; but here, we have to account for the fact that the frequency being counted is 9 MHz above the operating frequency, and for the fact that we have different offsets (carrier oscillator frequencies) for USB and LSB operation. This makes it necessary to preset the counters to a number different than zero in order to compensate. When operating in the USB mode, the number to which the counters must be preset are (MSD to LSD) 09967; in the LSB mode, it is 10000. This is accomplished by hard-wiring the USB and LSB lines to the appropriate load inputs of the counter chips, and using the rst line to drive the "load" control pins. Then, when the active counting portion of the count cycle begins again, each counter advances through zero (overflow) to a value appropriate for the transceiver operating frequency, and this value is displayed. Any carry from the MSD counter is discarded and ignored.

Another way of stating the matter is that the counters are preset to the above numbers, which correspond to a number that is 9,000.0 (or 9,003.3) subtracted from the overflow value of the counter chain, i.e., 10000.0. This is, in fact, a negative number, although the counter does not recognize it as such---in this way, the counter subtracts the carrier oscillator frequency from the L.O. frequency, as is required for a correct display.

It is not necessary for the 10 MHz digit to be arrived at by a counter, and to do so would add unnecessary complexity to the circuit. Instead, it is hard-wired such that, when the BAND switch is on the 160 through 40 meter positions, it is extinguished; when it is in the 20 meter position, Q109 is saturated and segments B and C of the display are grounded through appropriate current limiting resistors, and the numeral "1" is therefore displayed; and in the 10 and 15 meter positions of the BAND switch, Q108 is saturated, grounding the current limiting resistors for segments A, B, D, E and G, thus displaying the numeral "2".

The necessary control signals for the counter (clr, rst, and xfr), are generated by the time base/control circuitry. Crystal Y101, operating at a frequency of 1.31072 MHz, provides the stability necessary for accurate frequency measurement. C110 permits it to be adjusted to precisely the correct frequency. U114 contains the oscillator circuitry, as well as a divider with a factor of 2^{14}, or 16,384. The output of the counter is the square wave signal labelled "clock" on the timing chart. U115 is a decade counter, with outputs taken at the clock-divided-by-two (A) terminal, and the clock-divided-by-ten (D) terminal. The waveforms are shown in proper relationship in the timing diagram. The D output, inverted by Q105, is used as the clr signal; gate U103A generates the rst signal, and gate U103B generates the xfr signal.

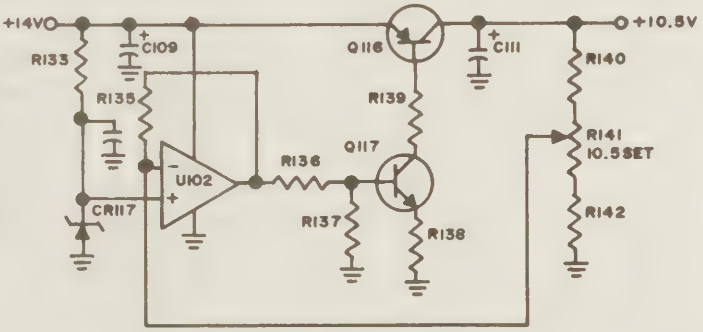
COUNTER TIME BASE



10.5 VOLT REGULATOR

A regulated 10.5 volts is required by many of the circuits in the transceiver. It is provided by the circuit comprised of Q116, Q117 and U102. The raw +14 volts is applied to the emitter of PNP series transistor Q116.

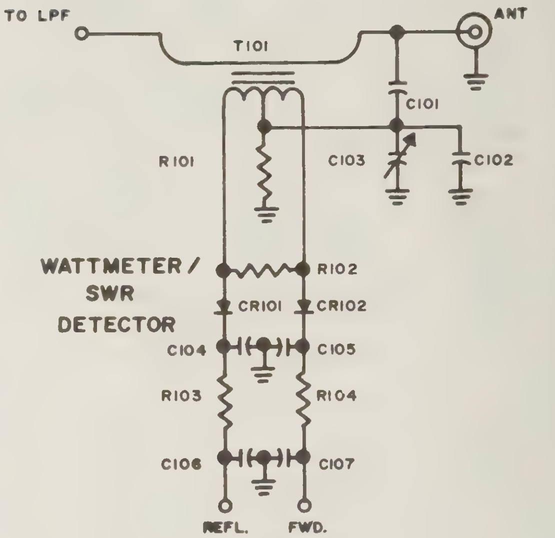
The output of Q116, its collector, is sampled by the R-140-R141-R142 voltage divider, with R141 serving as the "+10.5 set" control, and its wiper is connected to the inverting input of U102. A reference derived from zener diode CR117 through dropping resistor R133 is applied to the operational amplifier's non-inverting input. The output of this stage, a DC voltage, is applied to Q117 and determines the degree to which it is allowed to conduct, and thence its collector voltage. Its collector, in turn, drives the base of Q116 in such a manner as to control its degree of conduction, and thence to maintain its output voltage at the required level.



+10.5V REGULATOR

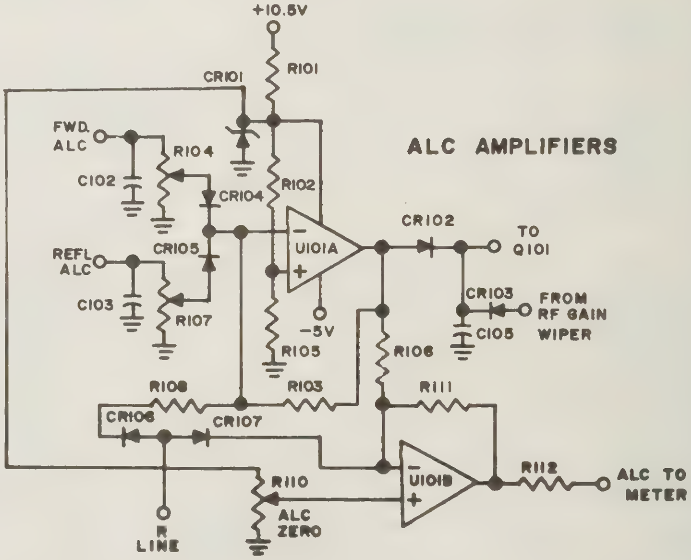
WATTMETER/SWR DETECTOR

The wattmeter board circuitry takes a sample of the current in the antenna line via transformer T101, and a sample of the voltage signal appearing on the antenna line at the output of the transformer via the capacitive voltage divider composed of C101, C102 and trimmer C103, the latter of which allows precise alignment. These signals are combined in such a manner that CR101 produces a rectified output voltage proportional to the magnetude of any signal reflected from the antenna due to mismatch, etc., while CR102 provides a voltage proportional to the power sent to the antenna. These signals are smoothed by the C104-R103-C106 and C105-R104-C107 combinations. These signals are then used in the ALC, SWR, and wattmeter circuitry to provide front panel indications and protective functions for the transmitter power amplifier.



AUTOMATIC LEVEL CONTROL

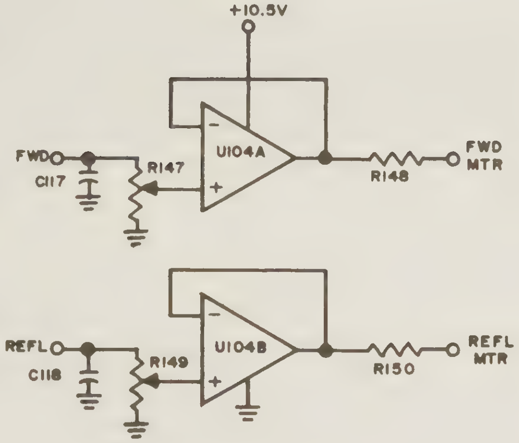
The FWD and REFL signals from the wattmeter board are coupled to the preselector board where they are adjusted by R104 and R107 respectively. C102 and C-103 bypass the lines as they come into the board. CR104 and CR105 act as an "or" gate, allowing the largest of the two signals to pass to the inverting input of ALC amplifier U101A. Feedback for this amplifier is via R103, while R102 and R105 establish a bias at the non-inverting input which sets the quiescent operating point of the output of this stage. R101 and CR101 provide a regulated voltage for this bias network and for U101, as well as for the ALC ZERO adjustment.



If the REFLected ALC signal should exceed a given amount, or if the FWD ALC signal should exceed that produced at the maximum safe output of the PA, CR105 or CR104 will conduct, causing the output of U101A to go more negative. This output will be passed through CR102, and on to Q101 in the preselector, where it is used to reduce the gain of the stage, and thus the output of the transmitter. Manual RF GAIN control is accomplished by inserting a voltage to Q101 via CR103 to the ALC line when in the receive condition. The "R" line, through CR106 and R108, biases the inverting input of U101A in such a manner as to disable it in receive.

The output of U101A is also coupled via R106 to the inverting input of U101B. The ratio of feedback resistor R111 to input resistor R106 determines the gain of the stage, and the output is coupled through limiting resistor R112 to the meter switch, where it may be selected by the operator to give a front panel ALC indication if he so desires. R110, connected to the non-inverting input of U101B, provides a zero set calibration control for the meter reading. When in the receive condition, U101B is also disabled by the "R" line, through CR107, in the same manner as U101A.

FWD and REFL signals from the wattmeter board are also applied to SWR meter drivers U104A and U104B through calibration potentiometers R147 and R149, respectively. These are unity-gain amplifiers, used only to prevent the meter from loading the wattmeter board output. The operator may select either the FWD or REFL power to be read on the calibrated meter by means of the front panel METER switch. R148 and R150 are current limiting resistors.

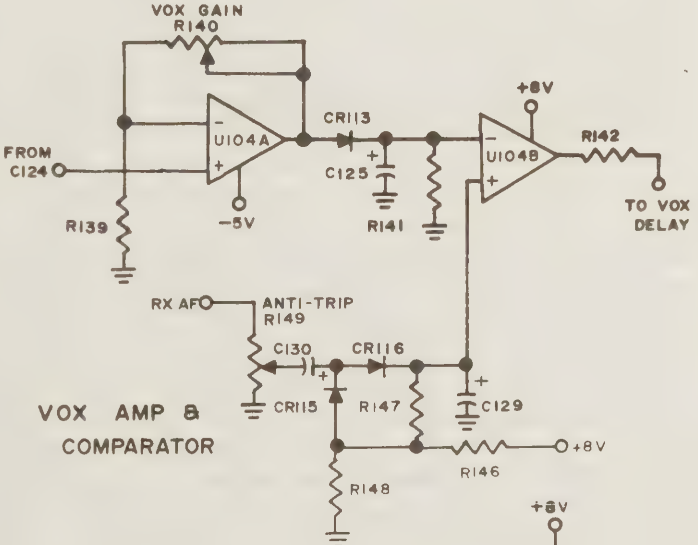


SWR METER DRIVERS

VOX CIRCUIT

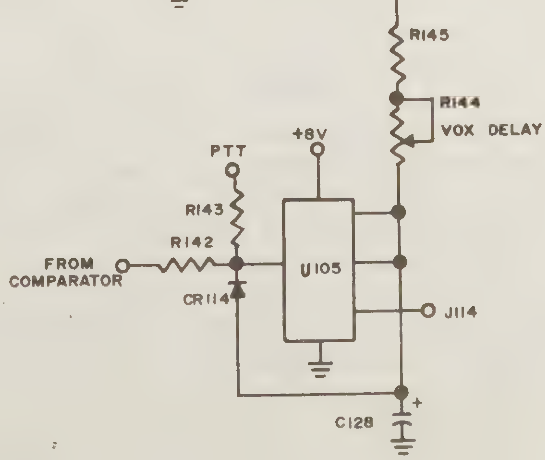
When VOX operation is selected by the operator, microphone AF is supplied from the preamplifier to U104A, where it is further amplified. The amount of this amplification is controlled by the VOX gain control, R140. The output of this stage is rectified by CR113, and smoothed by C125, and applied to the inverting input of comparator U104B.

A sample of the received AF, from the speaker line, is applied to the anti-trip control, R149, and then through C130 to CR115 and CR116, which form a voltage doubler rectifier. C129 smooths this DC signal, and it is coupled to the non-inverting input of the comparator. A DC bias, to set the comparator's threshold, is added to the rectified anti-trip signal by the R146-R148-network. The comparator's output will depend upon which of the two input signals are the larger; should the VOX signal be the largest, the output will go low; if the anti-trip signal is larger, it will be prevented from doing so.



VOX AMP & COMPARATOR

The output of the comparator is coupled through R142 to the trigger input of timer U105, causing it to activate and its output to go high. This high state is continued for the duration of the low on the input (diode CR-114 prevents C128 from charging during the duration of the low voltage at its cathode) plus the time it takes for C-128 to charge through R144 and R145 to 2/3 of the supply voltage after the low

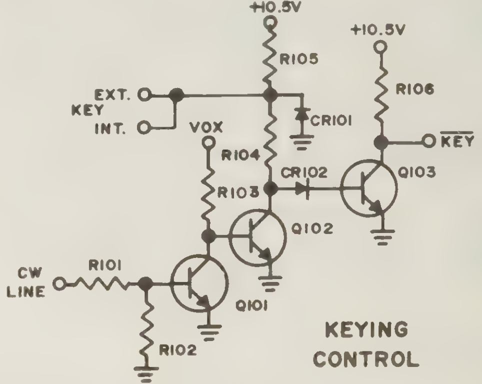


VOX DELAY

on the input of the timer has ceased. The amount of time it takes for C128 to charge is determined by the setting of R144, thus the amount of delay is controllable. Should VOX not be desired, the VOX/PTT switch connects a + voltage to the timer input via R143 when in the PTT position

KEYING CONTROL

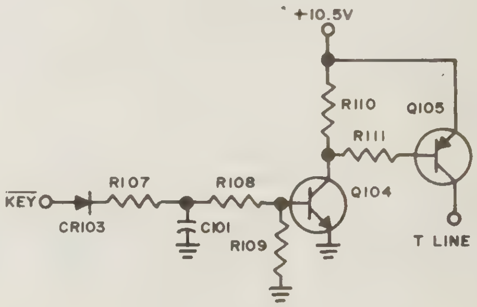
When in the CW mode of operation, the CW line is high, and it, through R101 biases Q101 on, effectively reducing its collector voltage and cutting off Q102. This makes the collector voltage of Q102 rise, and CR102 conducts in turn, biasing Q103 to an on condition. Thus its collector, the key line, is at a low value, its normal, un-keyed condition.



KEYING CONTROL

Should either the EXT KEY or the KEY contacts be closed, the junction of R104 and R105 in the collector circuit of Q102 is grounded, removing the voltage to CR-102 and the base of Q103 causing the collector voltage to Q103 (and the key line) to rise.

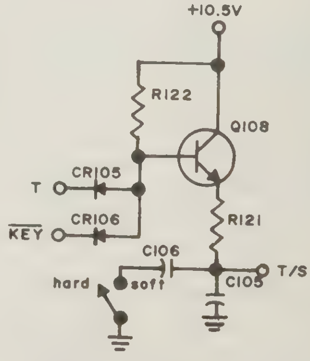
In the USB or LSB modes, the CW line is low, and Q101 does not conduct. PTT operation is then exactly as described above for CW, but if VOX is selected, the VOX circuit provides a voltage to R103 which causes Q102 to saturate, again removing voltage from CR102, and so Q103 ceases to conduct and the key line rises.



"T" LINE CONTROL

The key line, when high, causes CR103 to conduct through R107 and R108 biasing Q104 on. The combination of R107 and C101 control the rise characteristics of the keying waveform, while the falling characteristics are controlled by the discharge of C101 through R108 and the parallel combination of the base-emitter junction of Q104 and R109. When Q104 conducts, its collector goes low, lowering the base potential of PNP transistor Q105, and it saturates, energizing the "T" line.

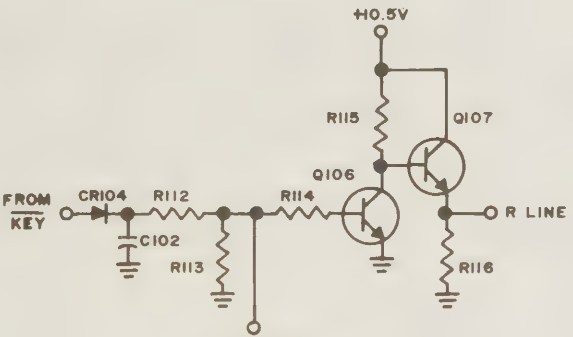
When the "T" line and key line are both high, CR105 and CR106 are reverse biased, allowing Q108 to conduct, and charge C105 through R121. This time constant determines how fast the "T/S" line is allowed to rise to full voltage. The "T/S" line supplies bias to the drivers for the transmit mixer, Z102 on the exciter board, thus permitting generation of the transmitter signal.



CW SHAPING

For lower speed CW operation, it is desirable to have a slower ("softer") attack and decay time on the CW note, so the operator may switch C106 in parallel with C105, thus increasing the time constant. High speed CW is more easily copied when it has a fast ("hard") attack time, so C106 is then switched out of the circuit, at the operator's option.

Also, when the key line is high, CR104 is allowed to conduct, charging C102 and biasing Q106 on through R112 and R114. This lowers the base voltage of Q107, causing it to cease to conduct, and bringing the "R" line to ground potential. When the key line goes low, C102 discharges through R112 and the R113-R114 and Q106 base-emitter junction parallel combination, delaying the turn off of Q106 and thus the return of the "R" line to a high value. This is to assure that there is a delay between the deenergizing of the "T" line and the reenergizing of the "R" line, so that certain switching functions can be accomplished without affecting either receive or transmit function.

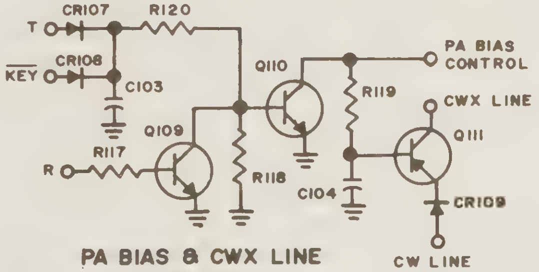


"R" LINE CONTROL

When semi-breakin is selected the CW timer holds the base of Q106 high for a brief time, preventing return to the receive condition between characters. When full breakin is selected, this feature is disabled.

When either the "T" line or the key line is high, voltage is applied to Q110, making it conduct. This brings the PA BIAS control line low, and turns on the bias to the power amplifier stages. Also, the base of Q111 goes low, allowing it to conduct. Q111 then conducts from the CW line through CR109, to the CWX line, energizing it.

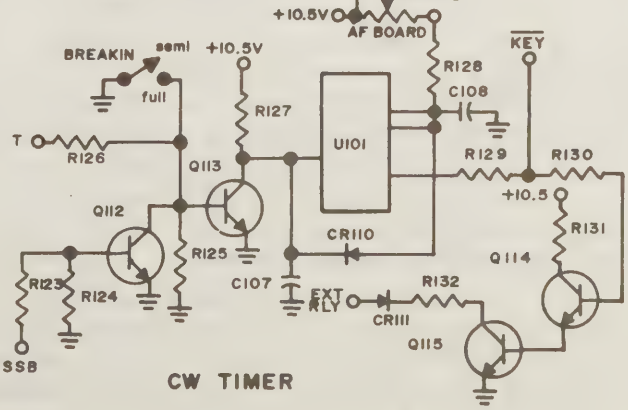
Should, however, the "R" line be high, voltage is applied through R117 to the base of Q109, causing it to saturate, and grounding the base of Q110, thus nullifying the effect of the "T" and key lines outlined above. Thus, there can be no output from the transmitter until the antenna switching circuitry has returned to the receive condition. This is an "interlock" function, which prevents damage to the receiver front end in the event of a control malfunction.



PA BIAS & CWX LINE CONTROL

CW TIMER

When the BREAKIN switch is in the SEMI position, energization of the "T" line applies a voltage to the base of Q113, causing it to saturate and the voltage at the input of timer U101 to go low. So long as this condition pertains, capacitor C108 is prevented from charging through R128 and the CW DELAY control by diode CR110. When the input is low, or until C108 is

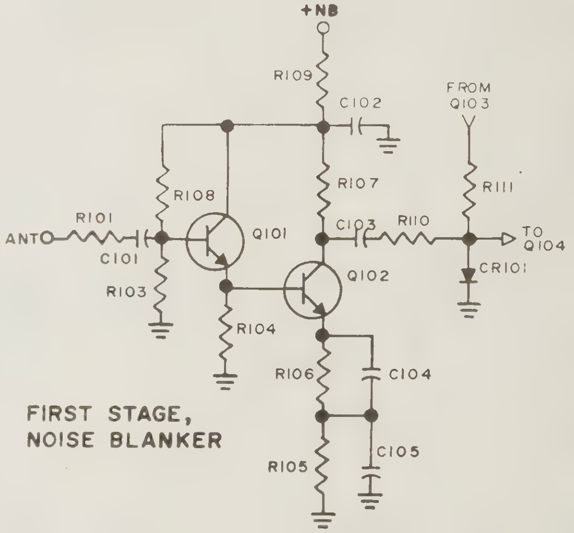


CW TIMER

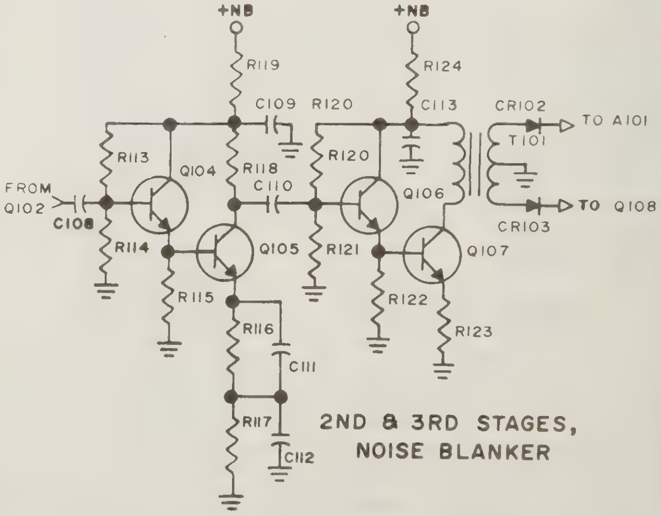
charged to 2/3 of the supply voltage after the input has returned high, the output of U101 is high. When this condition prevails, Q114 is turned on through R129 and R130. This in turn saturates Q115, which keys an external relay (such as for a linear amplifier, etc.) through protective diode CR111 and resistor R132.

Should FULL breakin be selected by the operator, the base of Q113 is grounded and the control input of the timer is never allowed to go low. Keying is then controlled by the $\overline{\text{key}}$ line, connected to the junction of R129 and R130. When the transmitter is keyed, this line goes high, and Q114 and Q115 are again saturated, grounding the external relay line. Note that few external amplifiers are capable of operating at full breakin, due to the slowness of the switching of the various relays, etc. Only those amplifiers especially built, or modified, for QSK operation should be used when full breakin is desired.

NOISE BLANKER



FIRST STAGE, NOISE BLANKER

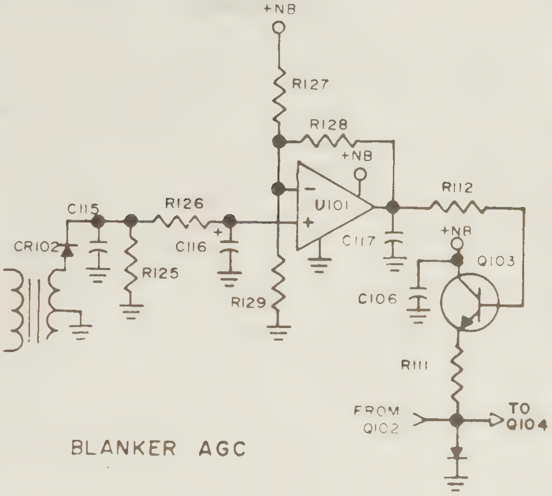


2ND & 3RD STAGES, NOISE BLANKER

The noise blanker signal input is taken directly from the receiver antenna input, and is coupled through isolation resistor-capacitor combination R101-C101 to the base of compound stage Q101 and Q102. Bias for that pair is derived from R108 and R103, with supply decoupling provided by R109 and C102.

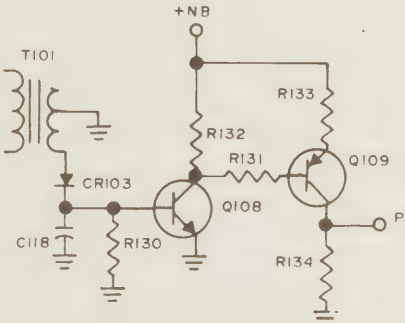
Frequency compensation is accomplished by R106-C104 and R105-C105 networks. Output is taken through C103, R110, and C108 to the base of the second compound stage. That stage, Q104 and Q105, operates in the same manner as the first one. In turn, this output is taken to yet a third compound stage, Q106 and Q107. This pair utilizes a conventional emitter network and untuned transformer T101 in its output. Note that the entire circuit is broadbanded--no selectivity is built in, so that noise pulses, no matter what frequency, will be accepted. The use of three compound stage amplifiers in cascade realizes a very high gain, for good sensitivity to the noise pulses.

The upper half of the secondary of T101 drives detector CR102, and after filtering by the C115-C116-R126 network, the resultant DC signal is applied to operational amplifier U101. The gain of this device is set by the ratio of R128 to R129 and its input threshold point is determined by voltage divider R127 and R129 on the non-inverting input. The DC output of A101 is applied through current limiting resistor R112 to the base of Q103. Q103 supplies a current to PIN diode CR101 which serves as a variable attenuator, thus supplying AGC to the noise blanker receiver. The R125, R126, C115 and C116 network integrates all incoming signals to drive U101 and generate AGC to prevent strong signals (SSB, AM, or CW) from generating cross modulation in the blanking gate.



BLANKER AGC

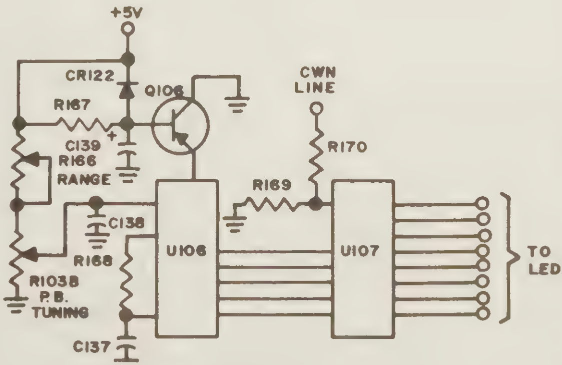
The bottom half of T101 drives detector CR103, which supplies a DC pulse to the base of Q108 when a noise pulse is received. When Q108 is thus turned on, it in turn causes PNP transistor Q109 to conduct, creating a blanking pulse across its load resistor, R134, which is coupled to the noise blanker gate in the I.F. circuit on the exciter board.



PULSE DETECTOR &
PULSE AMPLIFIER

PASSBAND TUNING INDICATOR

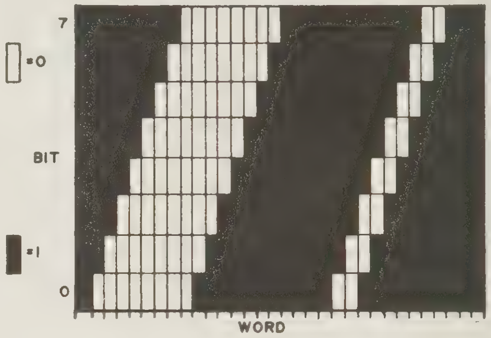
A separate section of the front-panel PASSBAND tuning control, R103B, is used to develop a signal used to drive the series of LEDs that indicate the status of the tuning of the passband circuit. R166 is used in series with it to provide a range calibration adjustment, and the output, bypassed by C138, is coupled to the analog input of analog-to-digital converter IC U106, on the audio board.



PASSBAND TUNING INDICATOR

U106 is capable of free-standing operation, but to operate in such a manner, it must be provided with an R-C combination that sets the frequency of operation of the internal clock oscillator on the chip. R168 and C137 serve this function. The IC must also be properly initialized when it is first powered up. Q106 and its associated circuitry perform this task. C139 charges through R167 when power is first applied, thus allowing Q106 to conduct until nearly full charge is achieved, grounding the appropriate pins on the chip. Once C139 has charged, Q109 ceases to conduct, and the initialization pins on the IC are ungrounded, allowing it to operate. When power is removed, CR122 allows C139 to discharge rapidly.

The digitally-coded output of U107 is coupled to the inputs of PROM U107. The internally stored code in the PROM generates the required signals to drive the front-panel LEDs used as the passband tuning status indicator. When in the CW narrow mode, the CWN line energizes another input of the PROM, changing the code presented to the LEDs such that only two adjacent devices, representing the frequency range being passed by the PBT circuitry, are lighted.



U107 MEMORY MAP

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Servicing Hints

Transistor Troubleshooting........................7-1
Integrated Circuit Troubleshooting..............7-2
Measuring S+N/N Ratio...........................7-2
Solid State Switching...........................7-3
Two Tone Testing................................7-3
Feedback Amplifiers.............................7-5
A Simple Impulse Noise Generator................7-6
The Phase Locked Loop Synthesizer...............7-7
Passband Tuning.................................7-8

CUBIC
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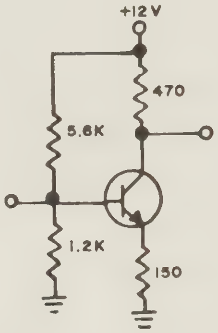
TRANSISTOR TROUBLESHOOTING

Many techniques for troubleshooting transistor circuits have been evolved, of varying degrees of complexity and reliability. One system, perhaps the simplest and most reliable, is outlined here for your information. With it, you can determine, with only a few minutes thought and basic arithmetic calculations, what the normal operating voltages on any given transistor stage should be, and by comparing them with those actually measured, locate the fault in the circuit.

The key to the scheme is to remember that, for normal operation, the base-emitter junction of a transistor must be biased into conduction--and that when this is so, the voltage drop across that junction will be approximately 0.7 volts for silicon devices and about 0.25 volts for germanium.



To illustrate how the system works, lets take a simple amplifier stage (at the right). To start with, we know that the base current is very small, on the order of microamperes. For our purposes, then, we can ignore it, for it would be masked by circuit tolerances sufficiently that we couldn't measure a voltage drop caused by it with our usual methods. We therefore treat the bias network (the 5.6 kΩ and 1.2 kΩ resistors) as a simple voltage divider, and can compute the voltage that would appear at their junction (the base of the transistor). In this case, it works out to about 2.1 volts. Since we know that this silicon transistor has a base-emitter drop of about 0.7 volts, it follows that the eimtter voltage is that amount less than the base voltage, or about 1.4 volts. The emitter current can then be calculated by Ohm's law, and comes out to be about 9.3 mA.

We know that both the base current and the collector current flow through the emitter resistor, but the base current is so small that we can say for practical purposes that the emitter and collector currents are the same; therefore, the current through the 470Ω collector resistor is also about 9.3 mA, and the voltage drop across it is therefore about 4.4 volts. This means that the collector voltage would be 12 - 4.4, or 7.6 volts.

Actual measured values on a normally operating circuit would be close to these values, while a malfunctioning transistor would result in greatly different values being found. Remember, transistors do not get "weak", as do vacuum tubes (with the exception of some emitter-ballasted power devices) so when they fail, they do so catastrophically--they open or short, or both. Thus, the voltages found on their terminals will be greatly different than if it were operating normally. For example, a shorted base-emitter junction would result in no 0.7 volt drop between those terminals; an open base-emitter junction would leave the emitter at ground potential (no base current flowing, so no collector current, so no drop across the emitter resistor). And so on. It isn't hard to figure out just what is wrong with the device, although this isn't really necessary--if the voltages are not close to what you calculated, the device is probably bad.

INTEGRATED CIRCUIT TROUBLESHOOTING

Very little can be written concerning the finding of internal faults in integrated circuits. Often, the exact internal circuitry is both complex and unknown to the service technician, for manufacturer's data sheets generally show only an equivalent circuit at best, and at worst, only the pinout.

The data sheets often do, however, give a truth table or timing (waveform) diagram for digital circuits, or some other description, not of what the internal circuitry _is_, but rather what it _does_. This, then, is the approach we must use in troubleshooting them also.

In short, we troubleshoot IC's by the process of elimination; Are the voltages at each of the pins about what they should be? Is the proper signal going into it? Then, if the proper signal isn't coming out of it, it is bad and must be replaced. Never mind what is wrong inside--you can't get inside to fix it anyway, and time spent figuring it out is wasted. Just replace it.

A good voltmeter, a high impedance VTVM or better yet one of the newer digital voltmeters that will read to tenths of a volt or less, and a good oscilliscope are then necessary pieces of test equipment, and no shop should be without them. If you don't have them, you'll have to spend your time in speculation, and for the most part, the results will depend on guesswork. Either you'll use more time than necessary (and have to charge the customer for it) or you'll wind up missing the cause of intermittant problems, replacing perfectly good parts, or the like. In either case, the increased cost to the customer (or decreased profit to the shop) is something that you can hardly justify.

MEASURING SIGNAL TO NOISE RATIO

The procedure for measuring the signal to noise ratio of a receiver is really quite simple, and it is a check that should be made on every receiver before it leaves your shop--finding a weak receiver and fixing the problem while it's in the shop is one of the best ways to keep a customer happy, and it will pay dividends in repeat business.

Connect an unmodulated signal generator that has a 50Ω output impedance to the antenna terminals, and tune it for about 1 kHz beat note audible in the speaker. Adjust the generator's output for about 0.25 uV at the antenna terminals, so that the AGC circuit isn't activated. (Note that some generators are calibrated in terms of their _open circuit_ output voltage, not what is present when connected to a load--check yours out, so you can get the right setting.)

Connect an AC voltmeter across the speaker terminals, and adjust the volume control for a convenient reading on the meter. Note that reading. Now remove the signal generator connection and short out the antenna terminals without changing anything else on the receiver. Note that reading.

The first is the reading for the signal-plus-noise; the latter is the reading for noise only. The S/N ratio is 20 times the common log of the ratio between the first and second values. Some AC meters have log (or dB) scales and can be used directly.

SOLID STATE SWITCHING

One of the more useful innovations in recent years is the application of solid state switching to signals within receivers and transmitters. It allows switching to be accomplished at the site, without the need for routing long signal leads to a bulky, complex, and often troublesome multi-pole mechanical switch. Instead, a simple switch controls a DC voltage which in turn is applied to all of the points where switching of the signal is to take place.

The simplest form of a solid state switch is the diode. We know that when it is reverse biased, it presents a very high resistance--high enough to be considered an open circuit when used in low or medium impedance circuits. Modern manufacturing techniques have reduced the junction capacitance of switching diodes so low that they can also be used effectively in most high impedance circuits, too. Thus, a diode which has a reverse bias voltage greater than the peak value of any AC signal applied to it, so that it does not conduct during any part of the signal cycle, will act as an open circuit.

Conversely, modern diodes have a very low "on" resistance, so if forward biased with a current greater than the peak signal current, so that it conducts during all parts of the signal cycle, it passes the signal with little or no attenuation.

All that is needed, then, is a circuit arrangement of capacitors to block the DC but pass the signal, (and perhaps bypasses, if high-level RF is near) and resistors to apply the bias and switching voltages, and the diode itself.

Field Effect Transistors (FET's) can be used as even better solid state switches. Remembering that the source to drain path of the FET is a pure resistance, normally conducting, we can see that, as with any other resistance, current can flow in either direction (bilaterally), from source to drain or from drain to source, depending on the polarity of the applied voltage. Therefore, AC signals will pass.

Now, if we apply a voltage to the gate that is sufficient to "pinch off" the source-drain path, the effective resistance of that path increases greatly, effectively opening the circuit, and thence the signal path.

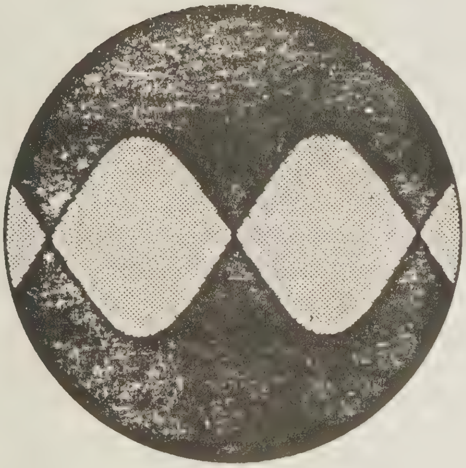
Normally, N-channel, depletion mode devices, with good bilateral characteristics in the source-drain path and low "on" resistance are selected for service in switching applications, although P-channel devices are also used.

All, then, that is needed for a FET switching circuit is the FET, suitable blocking capacitors and load and biasing resistors, just as in the case of the diode switch. In some instances, even these can be omitted, using the signal voltages themselves to provide any bias.

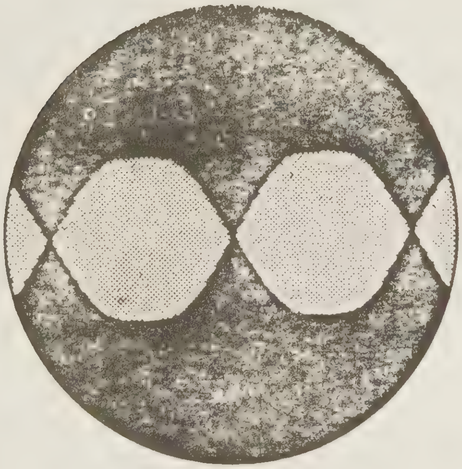
TWO TONE TESTING

Probably the most useful and yet least often performed test of a single sideband transmitter's performance a technician can make is the two tone test, and yet it is easily performed. The two tone test will disclose quickly and easily most of the major faults that are to be found in an operating transmitter by showing nonlinearities, improperly adjusted bias, overdriving, and many other problems.

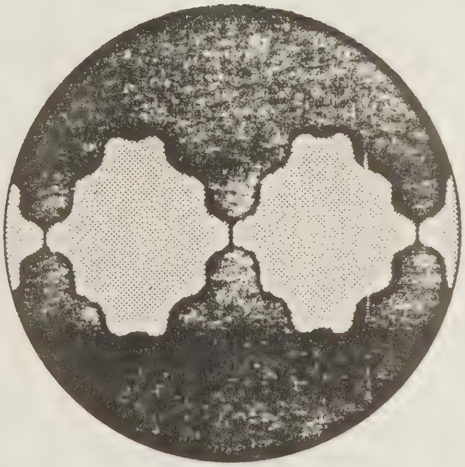
While a full discussion of the two tone test and its results is beyond the scope of this manual, a brief discussion is in order. More thorough



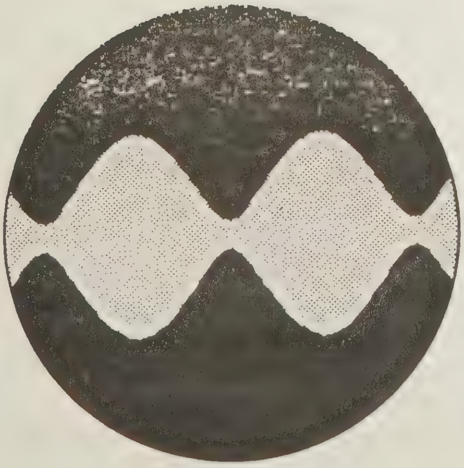
Two-Tone Pattern, Correctly Adjusted



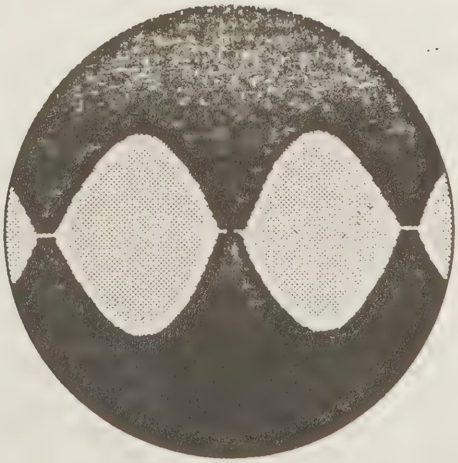
Two-Tone Pattern, Overdriven



Two-Tone Pattern, Carrier Present



Two-Tone Pattern, Unequal Tones



Two-Tone Pattern, Improper P.A. Bias

treatment is available in many competent references for those who wish to pursue the matter further.

Basically, this test is performed by applying two non-harmonically related audio tones, of equal amplitude and within the pass band of the radio, to the microphone input and observing the output across a dummy load on an oscilloscope. In this manner, the entire transmitter, from microphone jack to antenna terminal, is tested in one simple operation.

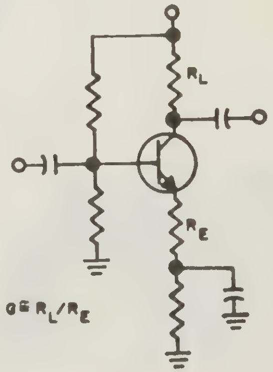
These tones can come from two audio signal generators, or better yet, from a two tone test set specifically designed for the purpose. Several such designs of adequate quality are included in amateur handbooks and other readily available literature. The tones must be very pure sine waves, and of equal amplitude, and not harmonically related to each other as mentioned.

The oscilloscope used to view the transmitter's output must be capable of displaying the frequencies at least as high as those in use, and preferably higher. A 30 MHz bandwidth instrument is adequate for all HF amateur bands, while the more common 15 MHz bandwidth instrument would serve for the 160, 80, 40 and 20 meter bands. Since any problems of nonlinearity, bias, etc., could be expected to show up on all bands, testing on one band only is usually considered adequate.

Several waveforms and their common causes are included here for reference, and others are available in current literature.

FEEDBACK AMPLIFIERS

The gain of individual transistors, within a given type number, unfortunately varies over a wide range. Often, this is satisfactory for the designer's purpose, but occasionally there arises a situation where the gain of the stage must be controlled more precisely. In that case, it is often advantageous to use an amplifier stage employing emitter degeneration (feedback) to set the gain of the stage. The accompanying diagram shows one such typical stage. The gain of the stage is approximately equal to the ratio of the collector load to the unbypassed portion of the emitter resistance. In the case of choke or tuned circuit loads, the value used in this calculation is the absolute value of the collector load impedance.

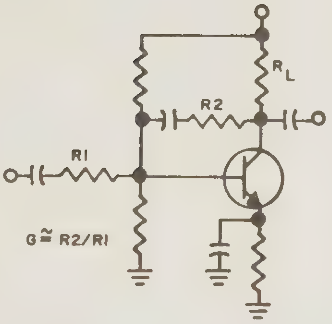


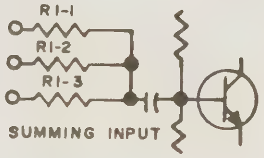
Utilizing this technique also raises the input and output impedances to higher values than would otherwise be the case, a condition that is often desirable. The bandwidth of the circuit can also be increased, under the right circumstances, although, under other conditions, it might be decreased instead. It is also possible to select the value of the capacitor bypassing a portion of the emitter resistance such as to provide frequency compensation for the inherent 6 dB per octave drop off in gain as the frequency is increased, that is a characteristic of all bipolar transistors.

Another type of feedback amplifier is shown in the second diagram. In this instance, the gain is approximately equal to R2 divided by R1, provided that the load resistance is much much greater than R2, and the impedance of the source feeding the circuit is much less than R1. The input impedance of this stage is equal to R1's value, and the output impedance is low. Since both input and output impedances can be quite low, the bandwidth may be very wide--100 MHz or greater.

Since the feedback path of the circuit (R2) allows just enough signal from the output to be fed back to the base to cancel out that coming in through R1 (and the feedback signal is out of phase with the incoming signal), little or no signal will be seen at the base. It is a "virtual" ground, so far as voltage is concerned!

This characteristic can be used to sum two or more signals, setting different gain values for each signal, if desired. The gain for each signal is proportional to the ratio of the feedback resistor (R2) to it's own input resistor only; the other input resistors do not affect the gain of that particular signal. It is interesting to note that this is a summing, not mixing, amplifier, often another useful property of the circuit.

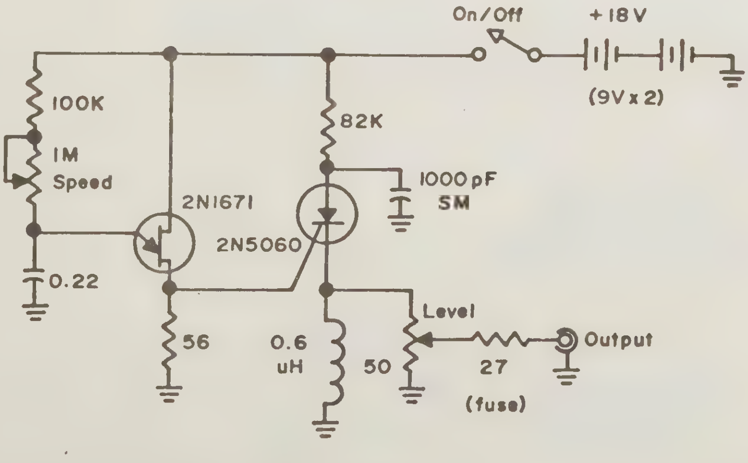




A SIMPLE IMPULSE NOISE GENERATOR

The noise blanker circuit used in SWAN radios, like all effective noise blankers, is designed to detect the presence of impulse noise above a certain threshold, and to mute, or blank, the output of the receiver for the duration of that pulse. (There is no circuit or scheme that can eliminate static or background hiss types of noise.) Since operation of the blanker circuit depends upon the presence of this high level impulse type of noise, it is difficult if not impossible to service the circuit unless a cource of such impulses is available. Efforts to service it using the commonly available test equipment lineup are ineffective, and may well lead to misadjustments--most often, this section of the receiver is never checked out, but merely accepted on faith.

It follows, then, that a simple, inexpensive instrument that generates this type of signal would be most useful in the service



Noise Generator

shop, yet commercial instruments are prohibitively expensive. For that reason, the circuit of such an instrument is included here, so that it may be built up locally and added to the shop equipment inventory for such service.

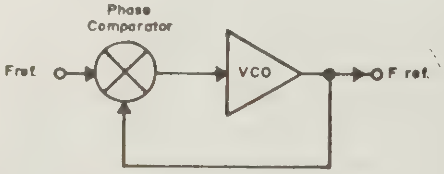
Referring to the accompanying diagram, we see that the 2N1671 is a unijunction transistor, connected in an oscillator circuit. The 0.22 mfd capacitor is charged through the 100K and 1M resistors, with the latter made adjustable to control the charging rate. When the charge reaches a certain level, the UJT fires, discharging the capacitor, and generating a sharp pulse of current through the 56 ohm resistor. This in turn fires the 2N5060 delivering thereby a relatively large current pulse through the inductor. This pulse is used to simulate the impulse noise. Level is adjusted by the 50 ohm potentiometer, and the small wattage 27 ohm series resistor acts as an output fuse, for protection from the circuit to which it is connected. Power is shown as two 9-volt batteries in series, although a suitable AC supply delivering approximately 18 volts DC at a few milliamperes would also be suitable.

THE PHASE LOCKED LOOP SYNTHESIZER

Electronic phase locked loops (PLL) have been in use since the 1930's where they were used for radar synchronization. In later years, they came into more widespread use as the synchronization system in television receivers and other applications. It is only in recent years that they have found widespread use as frequency synthesizers, where they have found many applications in satellite communications systems, airborne navigation systems, FM communications, and, most recently, in HF SSB communications equipment.

In the basic PLL system, the output of a local voltage controlled oscillator (VCO) is compared with an incoming, or reference, signal, and if the two are not of the same frequency and phase, an error signal is generated to tune the VCO in such a direction, and to the required degree, that they become so. This is the system used in television sync systems, and is shown in the accompanying diagram.

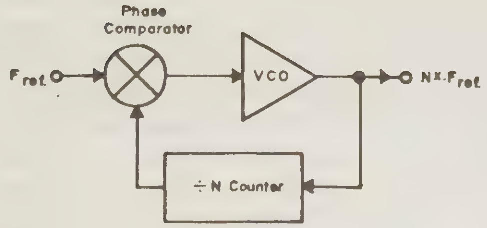
The usefulness of such a simple system is limited, however, to such synchronization tasks; however, it was found that, if a means of dividing the output frequency of the VCO could be inserted in the path between the VCO and the comparator, the output of the VCO could be made to automatically assume a frequency that was a multiple of the reference frequency, that multiple being equal to the division factor of the divider circuit.



BASIC PHASE LOCKED LOOP

By making the divider circuit programmable, that is, making it so that the division ratio could be varied at will, and making the VCO of an appropriate frequency range, true frequency synthesis could be achieved; that is, the frequency of the VCO could be made to vary, in integer multiple steps of the reference frequency. This scheme is shown on the following page.

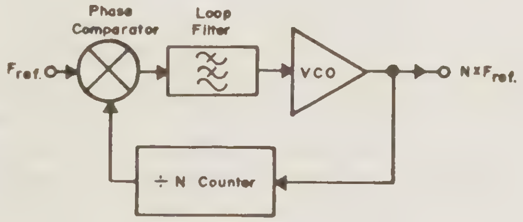
Digital circuitry lent itself quite handily to the implementation of the frequency divider, in that a number of available chips could be used as programmable dividers, or in the present terminology, "divide-by-N" counters, where N is the dividing integer. VCO's of various configurations were easily achieved, using varicaps as the tuning element.



BASIC MULTIPLIER PLL

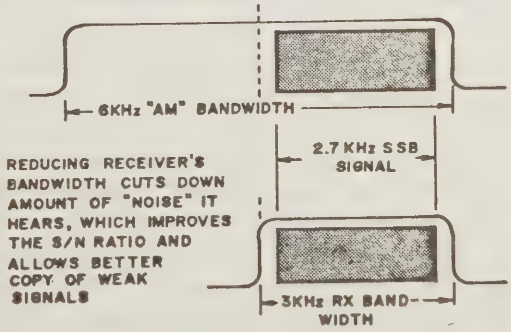
Initial PLL designs used phase comparators constructed out of discrete components, and were often very complex and difficult to "tame".

The development of specially designed integrated circuit phase comparators that were basically digital mixers greatly alleviated many problems, but created another. In such a mixer, leak-through of the reference frequency components to the error, or tuning line to the VCO is common unless specific steps are taken to prevent it. Such leak-through results in frequency modulation of the VCO at the reference frequency, and if held low enough, will be unnoticable in the reception of AM or FM signals, but is quite difficult to reduce to a level where SSB or CW signals can be received without objectionable distortion caused by this FMing. The means taken to reduce this distortion is the includion in the loop, between the phase comparator and the VCO, of a filtering arrangement to reduce the amplitude of the reference frequency feedthrough to the VCO. Usually a low pass filter is used for this purpose.



PRACTICAL MULTIPLIER PLL

Since the loop filter contains considerable capacity to ground, of necessity, and its cutoff frequency is below the reference frequency, it controls the time required for the loop to reachieve lock after a change has been introduced, such as when the divide ratio of the counter has been changed. At low reference frequencies, lockup time can become objectionably long. In the Astro 102BX, a synthesizer is used to generate the different band frequencies which are mixed with the PTO output to produce the local oscillator signal. The reference frequency is sufficiently high that a relatively simple loop filter is all that is required. Setting the division ratio into the programmable counter is accomplished by the BAND switch, in conjunction with a diode matrix and programmable read-only memory integrated circuit.

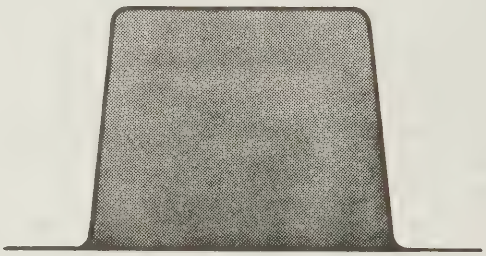

6KHz "AM" BANDWIDTH
2.7 KHz SSB SIGNAL
REDUCING RECEIVER'S BANDWIDTH CUTS DOWN AMOUNT OF "NOISE" IT HEARS, WHICH IMPROVES THE S/N RATIO AND ALLOWS BETTER COPY OF WEAK SIGNALS
3KHz RX BANDWIDTH

PASSBAND TUNING

It is a fact of communications life that the narrower the bandwidth of a receiving system, the more readily a signal of a given strength can be received and understood. This phenomenum occurs because what we really hear is the signal to noise ratio, rather than the absolute signal strength, and the narrower the received bandwidth, the less noise is received; hence a better signal to noise ratio and a more understandable recovered signal. The limiting factor, of course, is the bandwidth required for intelligibility of the desired signal. This is one of the chief advantages of single sideband over other communications methods. Since the necessary bandwidth is less than half that necessary for AM, for instance, it will result in more than twice the signal to noise ratio than an AM signal of comparable power, due solely to the reduction in received noise.

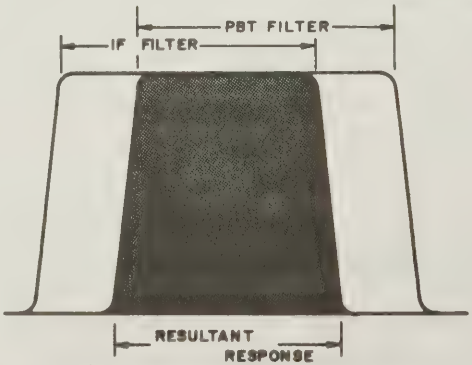
Studies show that the minimum bandwidth required for intelligible transmission of the human voice is about 2.5 kHz, although one can "get by" on a bandwidth of as low as 2.0 kHz, if a bit more degredation of the voice quality can be tolerated. In actual fact, however, these are average figures, and individual voice requirements vary widely. It has been the practice (and still is, in many cases), to use a fixed bandwidth, set by a high quality crystal filter in the IF of the receiver, to improve the signal to noise ratio. The response of a typical filter is shown to the right.

TYPICAL SSB FILTER CURVE



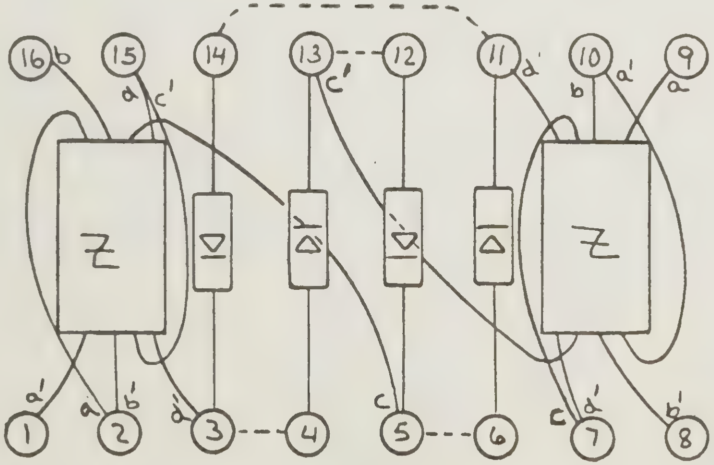
Since individual voices vary in their requirement for bandwidth, however, it would be advantageous in many instances, to be able to vary the bandwidth to lower limits. By being able to cut off the top or the bottom of the passband, thus reducing the bandwidth, it is possible to reduce the effects of close-by interfering signals, often at acceptable levels of degredation of the desired signal quality. Crystal filters, however, do not lend themselves to such operator adjustment.

Quite recently, a scheme for accomplishing this has been developed; it is called "Passband Tuning". In this scheme, the basic bandwidth is set by a high-quality crystal filter in the IF as before. But, after that filter, a circuit is included to translate that IF frequency to another frequency, usually higher. The signal is then passed through another crystal filter, of similar bandwidth as the first, and then translated back to the first IF frequency. If the frequency of the oscillator used to translate the signal to and back from the second IF is such that the passband of the two filters line up, no change in the bandwidth is noticed; however, if this oscillator frequency is made variable, the passbands of the two filters no longer lines up, and either the upper or the lower end of the passband of the first filter is cut off by the second, as shown in the accompanying diagram. Since the frequency of this passband tun-

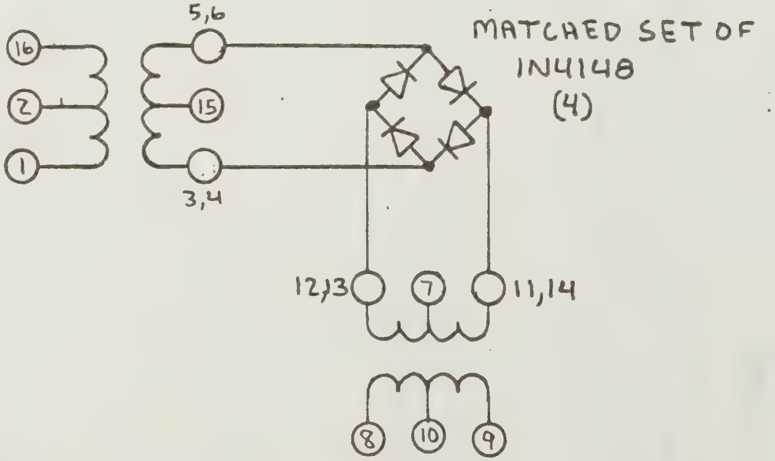


ing oscillator is continuously variable, we have achieved continuously variable bandwidth. Skillful operation of this system by the operator can result in useful copy of signals that would otherwise be uncopiable due to heavy interference from nearby signals. Such a scheme, with a unique visual indication of the bandwidth to which it is adjusted, is included in the Astro 102BX.

The following two pages provide data on the diode ring double balanced mixers used for Z2101, Z2102, and Z5101. The first type uses discrete diodes and the second uses a monolithic diode array. The two units are interchangeable.



TOP VIEW



5. THE FOLLOWING CONNECTIONS MADE EXTERNALLY
3 TO 4, 5 TO 6, 12 TO 13, 11 TO 14
4. MOUNTED ON 16 PIN SOCKET P.N. CH6170-16
3. WIRE COLORS
a - RED
b - BLUE
c - GREEN
d - CLEAR
2. USE (2) Z BEAD TRANSFORMERS, TYPE 1
1. USE (4) MATCHED DIODES 1N4148
NOTES

SWAN ELECTRONICS

SCALE: NONE
DATE: 5-26-77
APPROVED BY: Jcw
DRAWN BY C. DAVIDSON
REVISED

MIXER

DRAWING NUMBER
0800-2004

Servicing Data

Test Equipment Required..8-1
Initial Current and Light Checks...............................8-2
PTO Mode Switching and RIT Operation Check.....................8-2
PTO Range and Display Operation Check..........................8-3
Display Frequency Accuracy Check...............................8-3
Transmitter Operation Check....................................8-3
ALC Operation Check..8-4
Modulation Operation Check.....................................8-4
Receiver Operation Check.......................................8-5
Noise Blanker Operation Check..................................8-6
Passband Tuning Operation Check................................8-6
CWN Operation Check..8-7
PTO Operation Check..8-7
External Speaker Check...8-7
External L.O. Check..8-7

Alignment Procedure

- Initial Setup..8-8
- Voltage Regulator..8-8
- Band VCO...8-8
- Band Reference Oscillator....................................8-8
- Carrier Oscillator..8-9
- Passband Tuning Oscillator...................................8-9
- Counter Reference Oscillator.................................8-9
- Passband Tuning Indicator....................................8-9
- PTO and RIT..8-9
- Exciter I.F..8-10
- AGC and S Meter..8-10
- CWN Filter...8-10
- Notch Filter...8-10
- ALC and Meter..8-11
- Preselector and L.O. Bandpass Filters........................8-12

Level Diagrams

- L.O. Board...8-14
- Exciter Board..8-14
- Distribution Board...8-15
- Audio Board..8-15
- PTO Control Board..8-16
- Noise Blanker Board..8-16
- Counter Board..8-16
- Preselector Board..8-17

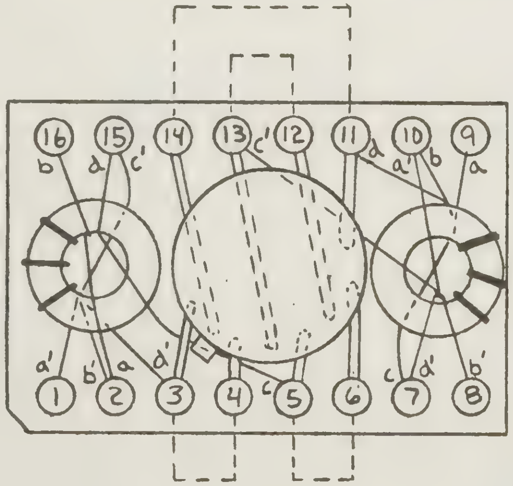
Band Synthesizer Memory Map...................................8-17
PBT Indicator Memory Map.......................................8-17

Layouts/Schematics

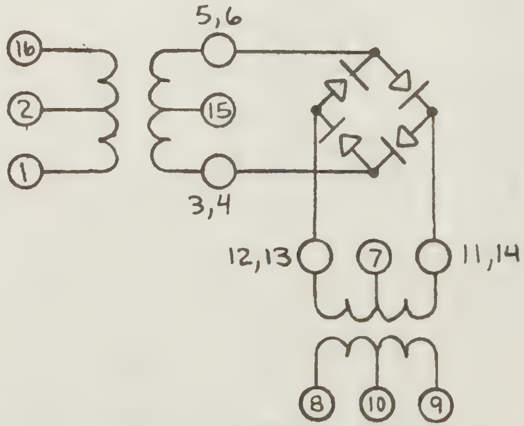
- PTO Control Board.................................8-18, 8-19
- Preselector Board.................................8-20, 8-21
- Low Pass Filter/Band Selector Board...............8-22, 8-23

REVISIONS

LTR	DESCRIPTION	DATE	APP'D
A	REVISED DWG. TO SHOW CA3039 DIODE PACKAGE INSTEAD OF SEPARATE DIODES CD 3-5-80	3-5-80	



TOP VIEW



4. (1) CA3039 MODIFIED PER DWG. NO. 0800-5404
3. (2) RF TRANSFORMERS 014-063
2. WIRE COLORS
   - a - RED
   - b - BLUE
   - c - GREEN
   - d - CLEAR
1. THE FOLLOWING CONNECTIONS MADE EXTERNALLY
   3 TO 4, 5 TO 6, 12 TO 13, 11 TO 14

NOTES

CUBIC COMMUNICATIONS, INC.
305 AIRPORT ROAD • OCEANSIDE, CA. 92054

SCALE: NONE	APPROVED BY:	DRAWN BY C. DAVIDSON
DATE: 3-4-80		REVISED

DOUBLE BALANCED MIXER

DRAWING NUMBER
0800-2004A

8½ X 11 PRINTED ON NO. 1000H-8 CLEARPRINT FADE-OUT

L.O. Board (Layout)....8-24
Counter/Display Boards....8-25, 8-26, 8-27
Audio Board....8-28, 8-29
Distribution Board....8-30, 8-31
Front Panel Right Board....8-32, 8-33
Passband Indicator Board....8-34, 8-35
Power Amplifier Board....8-36, 8-37
SWR/Wattmeter Board....8-38, 8-39
Noise Blanker Board....8-40, 8-41
PTO Board....8-42, 8-43
Exciter Board....8-44, 8-45
L.O. Board....8-46

Main Chassis/Interconnect Schematic....8-47

TEST EQUIPMENT REQUIRED

A certain amount of quality test equipment is necessary to service today's sophisticated communications eauipment. Below is a list of that required for the ASTRO 102BX.

- -Microphone, high impedance
- -Wattmeter, 0-200W scale
- -Dummy load, non-reactive, 50 ohms, capable of $\geq$200W dissipation
- -Light bulb load, 100W, 10 ohms cold
- -Oscilloscope, 30 MHz minimum bandwidth, with x10 attenuator probe to be used for all measurements
- -Frequency counter, at least 100 mV sensitivity, high impedance input, counts to 44 MHz or higher
- -VTVM or DVM with high input impedance to be used for all AC and DC measurements
- -Two-tone generator (See the "Servicing Hints" section of this manual)
- -Noise generator (See the "Servicing Hints" section of this manual)
- -Signal generator, 1.5-30 MHz, calibrated output
- -Audio generator, 300-3000 Hz
- -Power supply, 11-14V DC adjustable, 22 ampere or greater regulated output, with current metering
- -"T" coaxial connector
- -Attenuator for monitoring transmitter output with counter and oscilliscope

INITIAL INSPECTION AND SET-UP

Inspect the unit for physical damage; i.e., broken components, knobs, switches, meter, panel, heatsink, controls, circuit boards, cabinet, etc.

burned wires, components and boards; loose and improperly mounted plugs; frayed or pinched wires; loose hardware, etc.

Connect a suitable power supply to the POWER 13.8 volt connector on the rear apron of the unit. Connect the wattmeter, dummy load and "T" connector with the attenuator to the ANTENNA connector. Connect a suitable microphone to the KEY/MIC jack on the front panel.

Set the front panel controls as follows: AF GAIN off; RF GAIN, fully clockwise; IF GAIN fully clockwise; AGC DELAY fully counterclockwise: PASSBAND 12 o'clock; NOTCH FILT fully counterclockwise; MIC GAIN fully counterclockwise; PTO MODE to PTO A; MODE to LSB and BAND to 20.

INITIAL CURRENT AND LIGHT CHECKS

With the power off, the total current drain should be less than 15 ma. Turn the AF GAIN control to turn the power on, and the total current drain should be less than 2 amperes, the meter light should come on, the display should light, and the PTO A and PASSBAND KHZ lights should come on.

Key the unit with the microphone button. The total current drain should increase between 1 to 1.5 amps.

PTO MODE SWITCHING AND RIT OPERATION

Set the PTO MODE switch to EXT, and the display should read 11000.1 ± 1. Set the MODE to USB, and the display should read 10996.8 ± 1. Both PTO lights should be off.

Set the PTO MODE to A and adjust PTO A for 14260.0 with the RIT off (RIT light off). The PTO A light should be on and the PTO B light off. Key the unit. The display frequency should not change, and there should be no change in the PTO lights. Unkey the unit.

Set the PTO MODE switch to B and adjust PTO B for 14280.0 with the RIT off. The PTO B light should be on and the PTO A light should be off. Key the unit, and the display frequency should not change, with no change in the PTO lights. Unkey the unit.

Set the PTO MODE switch to REC A-XMIT B. The PTO A light should be on, the PTO B light should be off, and the display frequency should be 14260.0. Key the unit; the PTO A light should go off, the PTO B light should be on, and the display frequency should change to 14280.0. Unkey the unit.

Set the PTO MODE switch to REC B-XMIT A; the PTO A light should be off, the PTO B light on. The display frequency should be 14280.0. Key the unit, and the PTO A light should come on, and the PTO B light extinguish. The display frequency should be 14260.0 Unkey the unit.

Set the PTO MODE switch to PTO A, MODE to LSB. The display frequency should change less than 0.2 kHz. Set the RIT knob pointer to 0 and push it in to turn the RIT on. The RIT light should come on, and the display frequency should change less than 0.2 kHz. Adjust the RIT control from full counterclockwise to fully clockwise. The display frequency should vary between ± 0.8 to 1.5 kHz. Key the unit. The RIT light should go off, and the display frequency should be the same as if the RIT were off, and the RIT control should have no effect on the display frequency. Unkey the unit. Push the RIT control to turn it off. The RIT control should have no effect on the display frequency. Set the MODE switch to USB.

Set the PTO MODE switch to B and make the same checks as for PTO A.

PTO RANGE AND DISPLAY OPERATION

Set the PTO MODE switch to PTO A, turn the PTO A knob counter clockwise to its limit. The display frequency should be between 13940.0 and 13880.0. Turn the PTO A knob clockwise to its stop, and the display frequency should be between 14560.0 and 14620.0. Set the BAND switch to 10D. The display frequency should be stable. Set the BAND switch to 20.

Set the PTO MODE switch to PTO B and repeat the range checks for PTO B.

Set the BAND switch to 10B, and adjust PTO B for a display frequency of 28888.8 and check that the segments and decimal point are properly lighted on the display.

DISPLAY FREQUENCY ACCURACY

Connect a frequency counter to the "T" coupling and attenuator in the antenna line. Disconnect the microphone, and connect an audio generator to the EXT MOD jack on the rear apron of the unit. Connect an SPST switch to the KEY jack on the rear apron.

Adjust the audio generator's output to 1000 Hz ± 10 Hz. Key the unit with the SPST switch and adjust the MIC GAIN control and audio generator output level for approximately 20 watts output, or until the frequency counter has a stable count. Check the following:

That the carrier (output) frequency equals the display frequency minus 1.0 kHz in the LSB mode, ± 0.4 kHz.

That the carrier (output) frequency equals the display frequency plus 1.0 kHz in the USB mode, ± 0.4 kHz.

That the carrier (output) frequency equals the display frequency plus 0.8 kHz in the CW and CWN modes, ± 0.4 kHz.

TRANSMITTER OPERATION

Disconnect the frequency counter from the "T" connector in the antenna line and connect the oscilloscope in its place. Keep the attenuator in the line to protect the oscilloscope input circuitry. Disconnect the audio generator from the EXT MOD jack and connect the two-tone generator to the EXT MOD jack.

Check the power output and the total drain current on all bands in the CW mode, with the MIC GAIN control fully clockwise, and with the output frequency adjusted to the high end of each band. The power output should be 100 to 120 watts on all bands and the total current drain from the power supply should be 18 amps or less at 100 watts output. Power output should vary from less than 5 watts to full power as the MIC GAIN control is varied from fully counterclockwise to fully clockwise. There should be no output until the unit is keyed with the SPST switch.

Check the meter accuracy in the FWD position at 100 watts output on the 20 meter band. Accuracy should be ± 10%. Disconnect the coax at the ANTENNA connector on the rear apron, and set the METER switch to REFL, the MIC GAIN fully clockwise, and key the unit momentarily. The panel meter should indicate in the red area. Reconnect the coax to the ANTENNA connector. Set the meter switch to ALC, key the unit, and vary the MIC GAIN control from fully counterclockwise to fully clockwise. The meter reading should vary from approximately 0 to +20 or higher. Unkey the unit, and the meter should read approximately 0 in any of the METER switch positions.

ALC OPERATION

Disconnect the 50 ohm dummy load and substitute a standard 100W light bulb (10 ohms cold) as the load. Check the power output with the MIC GAIN control fully clockwise on all bands. Power output should be 40-70 watts if the ALC circuit is operating properly.

With the antenna connector open (i.e., no load), the total set current in transmit should not exceed 12 amperes. If so, the ALC circuit is not working properly.

NOTE: The purpose of the ALC is to limit power output and therefore PA device dissipation to safe limits, by limiting collector current. If there is too much ALC control, it will not be possible to obtain full output into loads with SWRs less than 1.5:1. If there is too little ALC control, then the PA transistors can draw excessive current, especially when the load is poorly matched, and possibly go into oscillation, causing them to overheat and fail. The temptation to increase the power output of the transmitter by reducing ALC control must be resisted--to do so will almost guarantee the failure of expensive components in the PA.

MODULATION OPERATION

Check the two-tone signal on each band and in both USB and LSB modes for crossover distortion, carrier leakage, and spurious outputs. None of the above should be detectable in the output (see the "Servicing Hints" section of this manual for examples and procedures for making the two-tone test.) Output should go to 100 W. PEP without distortion being introduced into the two-tone pattern.

NOTE: If a two-tone source is not available, a close examination of a modulated output display using a high quality microphone and/or monitoring the signal on another radio should show up any major discrepancy, although not so accurately as does the two-tone test.

Two-tone and microphone modulation checks should be made with the speech processor off. Disconnect the two-tone generator and reconnect the microphone. The MIC LEVEL trimpot (R138) on the audio board should be adjusted for 100 W. PEP output with the MIC GAIN control set at 10 o'clock and speaking into the microphone that is to be used with the unit, in a normal voice. Note the average power reading on the wattmeter. Switch the SPEECH PROCESSOR on and note that the average power reading increases by approximately 20 watts. It should also be noticable on the oscilloscope display that the peak to average modulation levels are less with the speech processor on than when it is off.

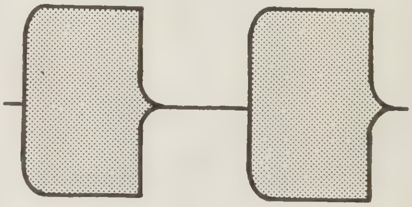
With the VOX-PTT switch in the VOX position, adjust the VOX GAIN trimpot (R140) on the audio board so that speaking into the microphone instantly keys the unit. Adjust the ANTI TRIP trimpot (R149) on the audio board so that audio from the speaker does not key the unit. Varying the VOX DELAY trimpot (R144) on the audio board should vary the unkeying delay time from 0 to 5 seconds. There should be no loud popping noises in the speaker as the unit keys or unkeys. Disconnect the microphone.

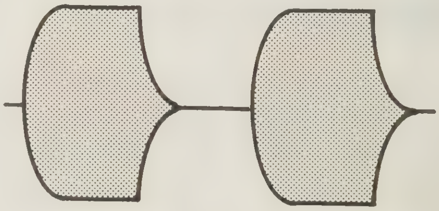
Set the MODE switch to CW, the BRK IN switch to FULL, the CW HARD-SOFT switch to HARD. Key and unkey the unit rapidly several times; the receiver audio should return the instant the unit is unkeyed. There should be no objectionable pops or clicks from the speaker when keying or unkeying at any power output setting. Set the BRK IN switch to SEMI. Key and unkey the unit several times. There should be a 0.5 to 6 second delay between unkeying and receiver turn on that varies with the setting of the CW DELAY trimpot (R174) on the audio board.

Set the BRK IN switch to FULL, key the unit, and adjust the MIC GAIN control for a reading of 75 on the meter with the METER switch in the ALC position. Rapidly key and unkey the unit and compare the oscilloscope display to the illustration presented here, to check for proper waveshaping. Set the CW HARD/SOFT switch to SOFT and make the same check for waveshape.

The SIDETONE trimpot (R124) on the audio board should vary the level of the audio tone heard in the speaker when the transmitter is keyed. The tone should be present only in the CW or CWN MODE of operation, and should not be heard with the unit unkeyed. Output should vary from no output to approximately 1 watt of audio with the setting of the sidetone level control.

Connect one lead of a 12 volt lamp (less than 200 ma) to +12 volts from the power supply and connect the other lead from the lamp to the center contact on the EXT RLY jack on the rear apron of the unit. The lamp should light only when the unit is in transmit in any mode.


"Hard" keying


"Soft" keying

RECEIVER OPERATION

Disconnect the microphone, the SPST switch, wattmeter, "T" connector and attenuator, and the dummy load from the unit. Connect an RF signal generator to the ANTENNA jack. Connect an AC voltmeter and frequency counter across the speaker terminals. Set the MODE switch to USB, VOX/PTT switch to PTT, MIC GAIN control fully counterclockwise, and PTO MODE to PTO A.

Check for 0.35 uV sensitivity at 10 dB S+N/N ratio on all bands. See the "Servicing Hints" section of this manual for the procedure. Check the sensitivity with the PTO MODE switch in the PTO B position also. Check that the sensitivity is better than 1.0 uV at 10 dB S+N/N ratio with the input voltage set at 11V DC on all bands.

Set the RF generator output to 1 uV and note the S meter reading. Tune the unit to each internally generated spurious signal and note that they each should be less than the S meter reading noted above.

Check that the S meter reads S 9 ± 1 S unit for 50 uV input signal at a supply voltage of 13.8 V DC. The S meter reading should vary with the input signal. Changing the input signal from 1 uV to 0.1 V should result in 6 dB or less increase in audio output (6 dB is a voltage ratio of 2). The

AF GAIN control should vary the audio output from inaudible to 3.5V RMS before clipping occurs, with 0.5 uV of input signal.

Turning the IF GAIN control fully counterclockwise should increase the S meter reading to full scale and decrease the units sensitivity. Set the IF GAIN control back to its fully clockwise setting.

Turning the RF GAIN control counterclockwise should decrease the units input sensitivity and the S meter reading should decrease. With the RF GAIN fully counterclockwise, input sensitivity should be reduced approximately 50 dB. Set the RF GAIN control back to its fully clockwise position.

With the AGC DECAY control fully counterclockwise, the unit should have no AGC delay time. This is indicated by the S meter reading increasing and decreasing instantly as the PTO is tuned across a signal. As the AGC DECAY control is advanced clockwise, the amount of time required for the AGC to decay should increase to approximately 15 seconds at fully clockwise. This is indicated by the slow drop in S meter readings after tuning off of a signal. Set the AGC DECAY control back to its fully counterclockwise position.

Adjust the RF generator so that the speaker audio tone is 1650 Hz ± 20 Hz. Adjust the NOTCH FILTER control for minimum speaker audio. The audio signal should decrease approximately 25 dB. Adjust the RF generator so that the speaker audio tone is 2500 Hz ± 20 Hz. Adjust the NOTCH FILT control for minimum speaker audio. The audio signal should decrease about 25 dB. Repeat the foregoing at 400 Hz ± 10 Hz audio tone. The audio signal should decrease by about 25 dB. Set the NOTCH FILT control back to its fully counterclockwise position.

NOISE BLANKER OPERATION

To check the noise blanker circuit, connect a pulse type of noise generator to the antenna connector along with the signal generator using the "T" connector. If a noise generator is not available, see the "Servicing Hints" section of this manual for the schematic diagram of a simple inexpensive set that may be constructed and kept for future use.

Tune in a 0.5 uV signal from the generator; turn on the noise generator and set the noise level to S 3 to S8 with the NOISE BLANKER switch off. Turn the NOISE BLANKER switch on; the signal should now be audible above the residual noise. Check on all bands.

Tune the signal generator to another frequency at least 1 MHz away from the set frequency and modulate the signal generator at 100% modulation. Increase the generator output to about 100,000 uV and at this point the noise level should come back up to S4 to S8; switching the NOISE BLANKER switch from on to off should have no effect.

PASSBAND TUNING OPERATION

With the PASSBAND control set at 12 o'clock, all eight PASSBAND KHZ lights should be on in LSB, USB, and CW modes. In CWN mode, only the 1.5 and 1.8 lights should be on. In LSB, USB and CW mode, as the PASSBAND control is turned clockwise, the PASSBAND KHZ lights should go off sequentially proceeding from left to right. At about the 4 o'clock position of the control, all lights should be out. As the PASSBAND control is turned counterclockwise from the 12 o'clock position, the PASSBAND KHZ lights should go off sequentially proceeding from right to left. At about the 8 o'clock position of the PASSBAND control, all lights should be out. Set the PASSBAND control back to the 12 o'clock position.

CWN TEST

Set the MODE switch to CW, and adjust the RF generator for 50 uV output and adjust the frequency for a 1650 ± 20 Hz speaker audio tone. Switch the MODE switch to CWN and note that the S meter reading changes by one S unit or less. Set the MODE switch to CW.

Adjust the RF generator frequency for a 2700 Hz ± 20 Hz speaker tone and set the RF generator output for S9 on the S meter. Turn the PASSBAND control counterclockwise until the S meter reading drops to less than S5. At this point, only the 2.7 and/or 2.4 PASSBAND KHZ lights should be off. Adjust the RF generator frequency for 400 Hz ± 20 Hz tone in the speaker and set the generator output for an S9 reading on the S meter. Turn the PASSBAND control clockwise until the S meter reading drops to less than S5. At this point only the 0.6 and/or 0.9 PASSBAND KHZ lights should be off. Set the PASSBAND control back to its 12 o'clock position.

PTO CHECK

Set the PTO MODE switch to PTO A and adjust the RF generator for 50 uV output. Tune PTO A across the signal and listen to the speaker audio tone for any indication of tuning backlash. Tune PTO A for a 1000 Hz ± 20 Hz tone in the speaker.

Set the PTO MODE switch to PTO B. Tune PTO B across the signal and listen to the speaker audio tone for any indication of tuning backlash. Tune PTO B for a 1000 Hz ± 20 Hz speaker tone. Both PTOs are now set to the same frequency and there should be no interference between the two. Slowly tune PTO A back and forth while listening for a change in the audio tone. There should be no detectable change.

EXTERNAL SPEAKER CHECK

Connect an external speaker to the EXT SPKR/PHONE jack on the rear apron of the set. The internal speaker should mute and the added speaker now should have audio output.

EXTERNAL L.O. CHECK

Disconnect the RF generator from the ANTENNA connector and reconnect it to the EXT LO jack on the rear apron of the unit. Adjust the RF generator frequency to 30 MHz and approximately 0 dBm output. Set the MODE switch to LSB, PTO MODE switch to EXT, and BAND to 15. The display frequency should be the RF generator output frequency minus 9000.0 kHz ± 0.1 kHz. The display should be stable and tuning PTO A or PTO B should have no effect on the display frequency.

This completes the operational test of the unit. Disconnect all test equipment and reinstall any covers that may have been removed.

ALIGNMENT PROCEDURE

Alignment should be performed only after making a complete performance check of the unit. It is advisable to then align only those areas of the unit necessary to bring performance up to specification. Also, the replacement of a defective component will usually restore the units performance without requiring any realignment, in most cases. Following this procedure will save time and help to avoid the possibility of inadverdantly misaligning a section of the unit.

INITIAL SETUP

Remove the top and bottom covers of the unit. Set the front panel controls as ollows: AF GAIN to off; RF GAIN to fully clockwise; IF GAIN to fully clockwise; AGC DECAY to 12 o'clock; PASSBAND to 12 o'clock; NOTCH FILT to fully counterclockwise; MIC GAIN to fully counterclockwise; PTO MODE to PTO A; MODE to LSB; BAND to 160 and PTT/VOX switch to PTT.

Connect a suitable power supply to the POWER 13.8 V connector on the rear apron of the unit.

VOLTAGE REGULATOR ADJUSTMENT

Turn the AF GAIN control clockwise to turn the power on. Adjust the power supply voltage for 14 V DC.

Measure the voltage at the collector of Q116 on the distribution board and adjust the 10.5 V SET trimpot (R141) on the same board for a reading of 10.55 ± 0.05 V DC on the voltmeter. Vary the power supply voltage from 11 to 15 volts. The measured voltage should change by less than 0.2 V DC. Reset the power supply voltage to 14 V DC.

BAND VCO ADJUSTMENT

Measure the voltage at the point labeled 160-15 M EL (near C138) on the L.O. board. Adjust the coils according to the following chart for 3.75 ±0.25 V DC.

BAND	COIL
160M	T101
80M	T102
40M	T103
20M	T104
15M	T105

Measure the voltage at the point labeled 10 M EL on the LO board. Set the BAND switch to 10A and adjust T106 for 3.20 ± 0.1 V DC. Switch to 10B, 10C, and 10D and note that voltage reading increases approximately 0.4 V DC for each band.

BAND REFERENCE OSCILLATOR ADJUSTMENT

Connect a frequency counter to collector of Q111 on the LO board. Set bandswitch to 10D. Adjust the 2 MHz oscillator trimmer capacitor (C246) for a frequency of 44.000 MHz ±20 Hz. Disconnect the frequency counter.

CARRIER OSCILLATOR ADJUSTMENT

Connect the microphone to the MIC/KEY jack on the front panel. Connect the oscilloscope probe to the emitter of Q117 on the exciter board. Key the unit with the microphone button and adjust transformer T105 on the exciter board for a maximum indication on the oscilloscope of approximately 1 V p-p. Unkey the unit. Disconnect the oscilliscope probe and reconnect the frequency counter to the same point. Key the unit and make adjustments according to the following table, in the order listed.

MODE	CAPACITOR	FREQUENCY
LSB	C194	9.00000 MHz ±10 Hz
USB	C197	9.00330 MHz ±10 Hz
CW	C199	9.00250 MHz ±10 Hz

Unkey the unit and disconnect the frequency counter and microphone.

PASSBAND TUNING OSCILLATOR ADJUSTMENT

Measure the voltage at the center leg of the PASSBAND tuning control (R103B, rear section) and adjust the PASSBAND control for 5.25 ± 0.05 V DC.

Connect the frequency counter to G2 of Q113 on the exciter board and adjust the PBT trimmer capacitor (C206) for a frequency of 22.80165 MHz ± 20 Hz. Vary the PASSBAND control from fully counterclockwise to fully clockwise and note that the PBT oscillator has a frequency range greater than 3.6 kHz.

Disconnect the frequency counter and reconnect the oscilloscope probe to the same point. Adjust coil T106 for maximum indication on the oscilloscope screen, approximately 1.5 V p-p. Disconnect the oscilloscope probe.

COUNTER REFERENCE OSCILLATOR ADJUSTMENT

Remove the cover from the units frequency counter and connect the shop frequency counter to pin 9 of U114 on the counter board. Adjust the reference oscillator trimmer (C110) for a frequency of 1.31072 MHz ± 2 Hz. Disconnect the frequency counter and reinstall the counter cover.

PASSBAND TUNING INDICATOR ADJUSTMENT

Set the PASSBAND control to its 12 o'clock position and adjust the A-D RANGE trimpot (R166) on the audio board so that all eight PASSBAND KHZ lights are on.

PTO AND RIT ADJUSTMENT

Set the MODE switch to LSB, and the BAND switch to 20, turn the RIT to its off position (RIT light extinguished) and adjust PTO A for a display frequency of 14270.0. Switch the MODE switch to USB, and if the display frequency changes, adjust the USB SHIFT trimmer (C608) on PTO A so that there is no change in the display frequency when shifting between LSB and USB modes.

Set the RIT control to 0. Push the RIT control to turn it on (RIT light lighted). If the display frequency changes, adjust the RIT CENTER trimpot (R536) on the PTO control board so that there is no change in the display frequency between RIT on and RIT off. Verify that the RIT control varies the frequency of the display ± 800 to 1500 Hz.

Turn the PTO A control fully counterclockwise and note how much the display frequency goes below 14000.0 kHz. Now turn the PTO A control to its clockwise limit and note how much the display frequency goes above 14500.0 kHz. The amount that the frequency extends beyond the lower and upper band limits should be greater than 50 kHz and within 10 kHz of each other. If this is not the case, adjust trimmer C606 on PTO A for the proper frequency range.

Set the PTO MODE switch to PTO B and repeat the above procedures for B.

EXCITER I.F. ADJUSTMENTS

Set the MODE switch to CW, the PTO MODE switch to PTO A, the PASSBAND control to its 12 o'clock position, and adjust PTO A for a display frequency of 14300.0 kHz. Connect the output of the RF generator to the ANTENNA connector on the rear apron of the unit. Set the RF generator output for 50 uV and adjust its frequency for maximum S meter reading on the unit. Adjust transformers T101, T102, T103 and T104 on the exciter board for a maximum S meter reading.

AGC AND S METER CALIBRATION

Use the same test setup as previous section. Connect the oscilloscope probe to the cathode of diode CR123 (AF in line) on the audio board. Adjust the AGC THRESH trimpot (R101) on the audio board for 600 mV p-p indication on the oscilliscope screen. Disconnect the oscilloscope probe.

Adjust the S METER CAL trimpot (R110) on the audio board for an S 9 indication on the unit's meter.

CWN FILTER ADJUSTMENT

Set the MODE switch to CWN and connect the frequency counter across the units speaker terminals, keeping the rest of the test setup the same as for the previous section. Adjust PTO A for a 1650 Hz ± 20 Hz speaker audio tone and adjust CWN trimmer capacitors C130 and C134 on the exciter board for a maximum S meter reading. Switch the MODE to CW and verify that the S meter reading changes by less than one S unit.

NOTCH FILTER ADJUSTMENT

Use the same test setup as before. Connect the AC voltmeter or oscilloscope across the speaker terminals. Adjust PTO A for a 1650 ± 20 Hz tone in the speaker. Adjust the NOTCH FILT control for minimum speaker audio output, then adjust the NOTCH FILTER ADJUST trimpot (R178) on the audio board for minimum speaker audio output. Repeat the adjustments of NOTCH FILT and N F ADJ controls until there is no further decrease in the speaker audio output. The notch filter should attenuate the speaker audio by more than 25 dB. Verify that speaker audio output can also be attenuated by the notch filter with speaker audio tones of 400 Hz and 2500 Hz. Disconnect the frequency counter, AC voltmeter or oscilloscope and RF generator from the unit.

ALC AND METER ADJUSTMENTS

Connect the microphone to the MIC/KEY jack on the front panel of the unit. Connect the RF wattmeter and 50 ohm dummy load to the ANTENNA connection on the rear apron. Set the MODE switch to CW and adjust PTO A for a display frequency of 14500.0 kHz.

Connect the DC oscilloscope probe to the "high" potential leg of the REFL trimpot (R107) on the preselector board. Key the unit with the microphone button and adjust the MIC GAIN control for 100 watts of output. Quickly adjust the SWR NULL trimmer capacitor (C103) on the SWR/wattmeter board for minimum DC voltage on the oscilloscope (< 0.5 V DC).

NOTE: During this and subsequent CW tests, limit the "key down" time to 30 seconds or less to prevent overheating and resultant damage to the power amplifier devices. Allow time between transmissions to permit them to cool.

Disconnect the oscilloscope probe. Set the METER switch to the FWD position. Key the unit, and adjust the MIC GAIN for 100 watts of output and quickly adjust the FWD trimpot (R147) on the distribution board for a reading of 100 on the unit's meter.

Set the MIC GAIN control fully clockwise, key the unit, and quickly adjust the FWD ALC trimpot (R104) on the preselector board for 110 watts of output.

Disconnect the 50 ohm dummy load and reconnect the 100 watt (10 ohm cold) light bulb in its place. With the MIC GAIN control fully clockwise, key the unit and quickly adjust the REFL ALC trimpot (R107) on the preselector board for 50-60 watts of output. Set the METER switch to REFL, key the unit, and quickly adjust the REFL trimpot (R149) on the distribution board for a reading of 100 on the unit's meter. (10 watts reflected power)

Check the output on all bands for 40 to 70 watts of output using the light bulb as a load.

Disconnect the light bulb load and reconnect the 50 ohm dummy load. Check the power output for 100-120 watts on all bands, with the MIC GAIN control fully clockwise.

Disconnect the coax lead at the ANTENNA connector of the unit. Key the unit with the MIC GAIN control fully clockwise and note that the total current drawn from the power supply is less than 12 amperes on all bands. Reconnect the coax lead to the ANTENNA connector.

NOTE: It may be necessary to slightly readjust the FWD and REFL ALC trimpots and SWR NULL trimmer capacitor to obtain the proper indications on all bands.

Set the BAND switch to 20 meters, the METER switch to ALC and the MIC GAIN control fully counterclockwise. Key the unit and adjust the ALC ZERO trimpot (R110) on the preselector board for a reading of S 2 on the units meter. Adjust the MIC GAIN control to its fully clockwise position. The meter reading should increase to +20 or greater. Unkey the unit.

PRESELECTOR AND L.O. BANDPASS FILTER ADJUSTMENT

The alignment of the bandpass filters should be performed only if absolutely necessary, and then only for those bands that require alignment. Indications of misalignment are significant reduction in power output and receiver sensitivity that is not the result of a component failure or misalignment of transformers, coils and/or oscillators on the exciter board.

If alignment is to be performed using a sweep generator, follow the procedure supplied with the test equipment and test the boards as follows:

Preselector Bandpass Filters: (with sweeper)

Connect the sweep generator output to the RX IN jack (J132) and the spectrum analyzer's input to the RX OUT jack (J136). Sweep each band according to the accompanying chart.

L.O. Bandpass Filters: (with sweeper)

Disconnect the leg of C177 from pin 7 of Z101 on the LO board and connect the sweep generator output to the leg of C177. Connect the analyzer input to the LO OUT jack (J103).

Sweep each band according to the accompanying chart.

Preselector Bandpass Filters: (without sweeper)

Disconnect the TX OUT coax at J134 on the preselector board and connect the oscilliscope probe to J134. Set the MODE switch to CW, and MIC GAIN control to fully clockwise, and key the unit. Adjust the oscilloscope for a reasonable indication.

Adjust the bandpass coils for maximum indication on the oscilloscope at the center frequency of each band.

Tune the PTO across each band and readjust the bandpass coils as necessary to obtain a flat response (± 2 dB) across the band without significantly reducing the maximum output.

L.O. Bandpass Filters: (without sweeper)

Disconnect the LO OUT coax at J103 on the LO board and connect the oscilliscope probe to J103. Use the same procedure for adjusting these bandpass coils as in the preceeding section.

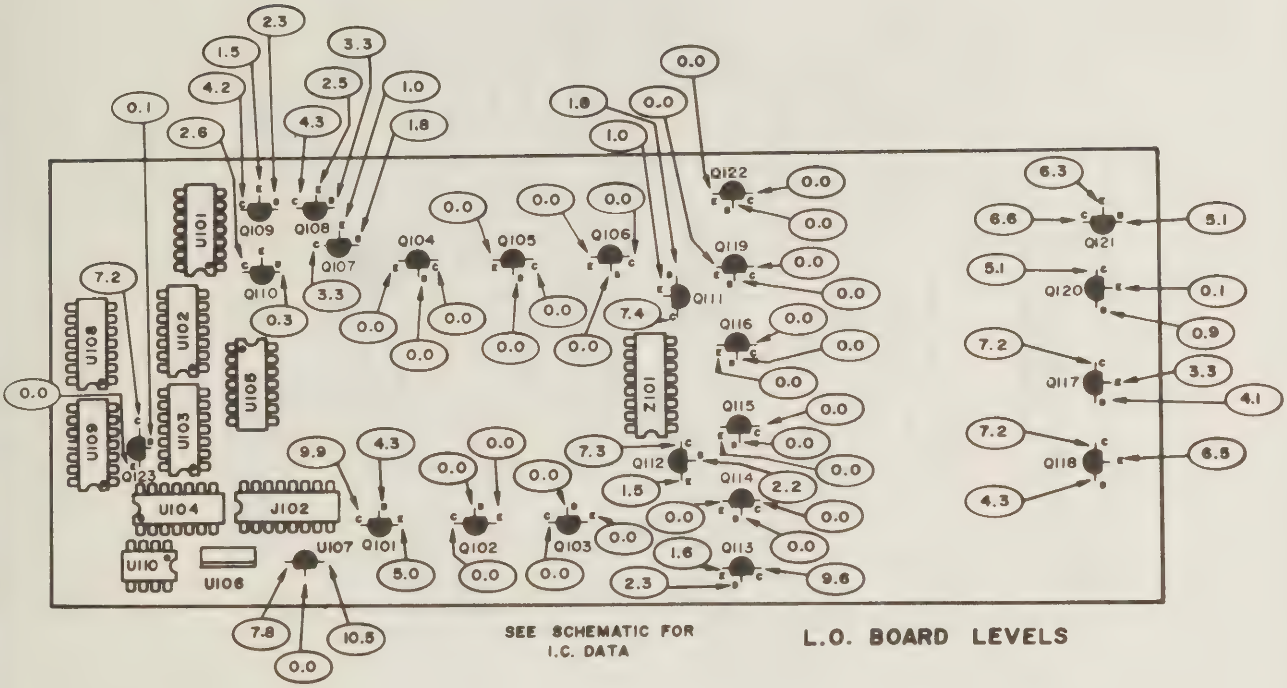
NOTE: It is not necessary to key the unit for LO bandpass filter adjustment.

PRESELECTOR BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHZ	CENTER FREQ.	COILS
160	1.8-2.0	1.9	T101, T102, T103
80	3.5-4.0	3.750	T104, T105, T106
40	7.0-7.5	7.250	T107, T108, T109
20	14.0-21.5	14.250	T110, T111, T112
15	21.0-11.5	21.250	T113, T114, T115
10A, B, C, D	28.0-30.0	29.0	T116, T117, T118

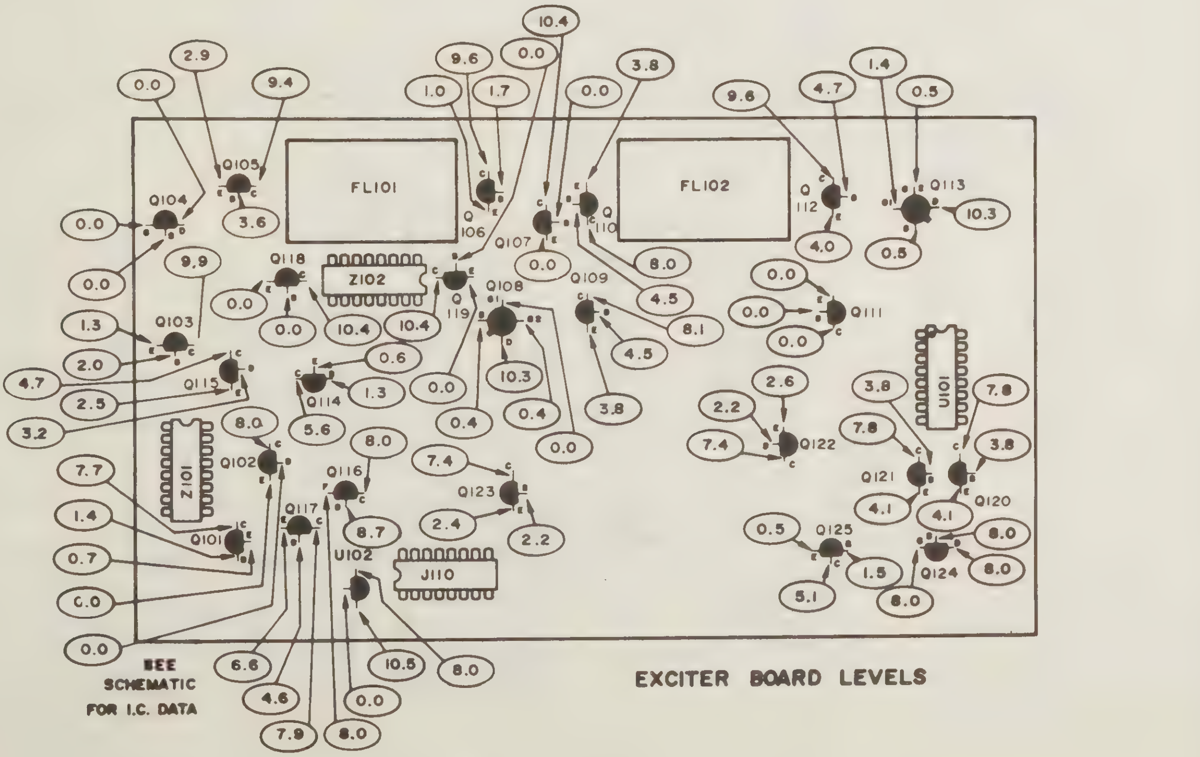
L.O. BANDPASS FILTER ALIGNMENT TABLE

BAND	RANGE, MHZ	CENTER FREQ.	COILS
160	10.8-11.0	10.9	T107, T108, T109
80	12.5-13.0	12.750	T110, T111, T112
40	16.0-16.5	16.250	T113, T114, T115
20	23.0-23.5	23.250	T116, T117, T118
15	30.0-30.5	30.250	T119, T120, T121
10A, , C, D	37.0-39.0	38.0	T122, T123, T124



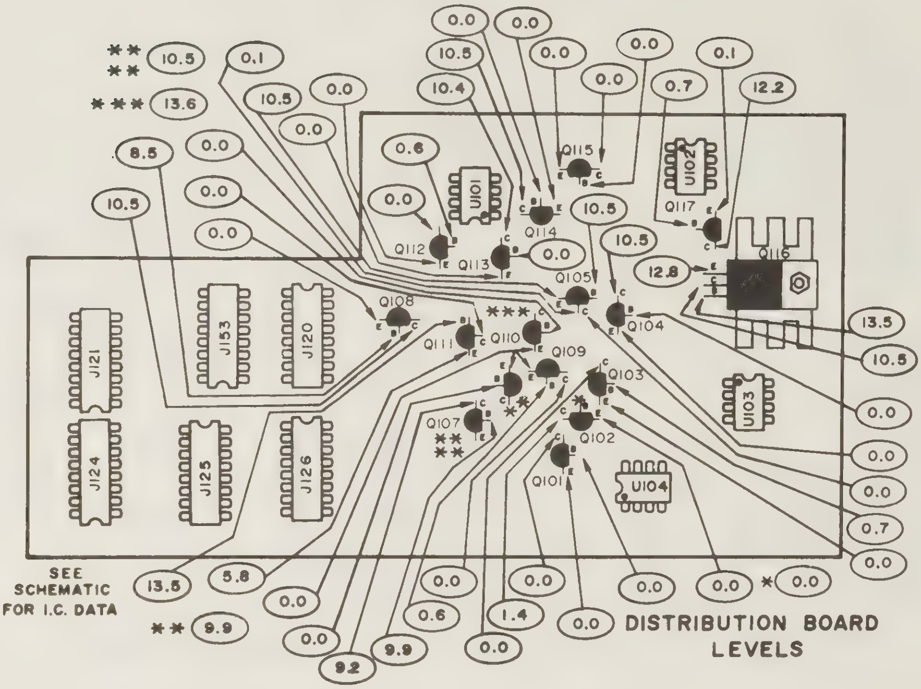
SEE SCHEMATIC FOR I.C. DATA

L.O. BOARD LEVELS



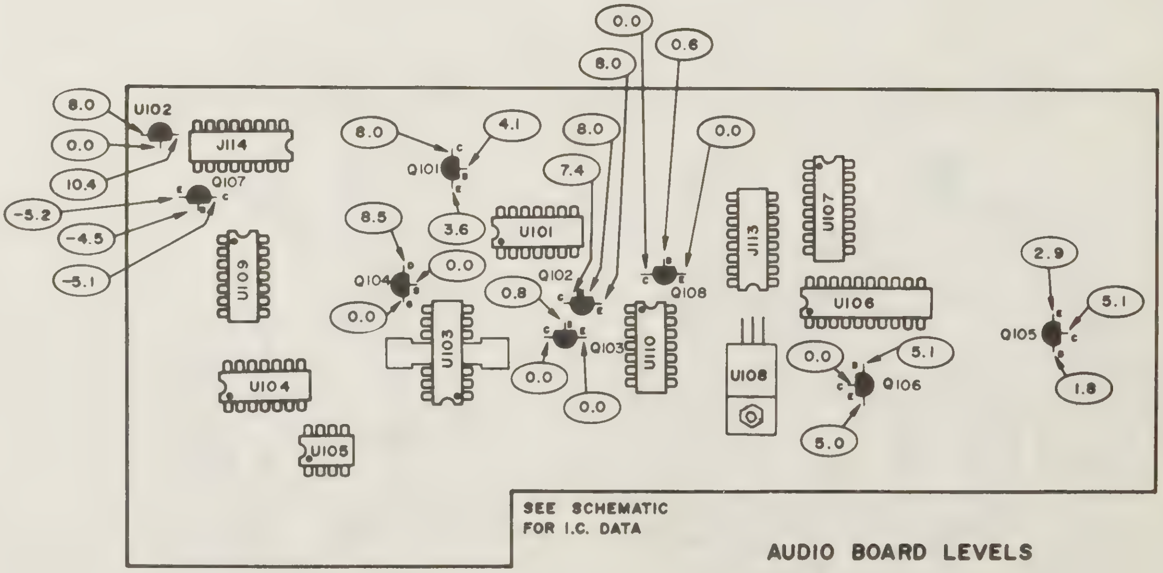
SEE SCHEMATIC FOR I.C. DATA

EXCITER BOARD LEVELS



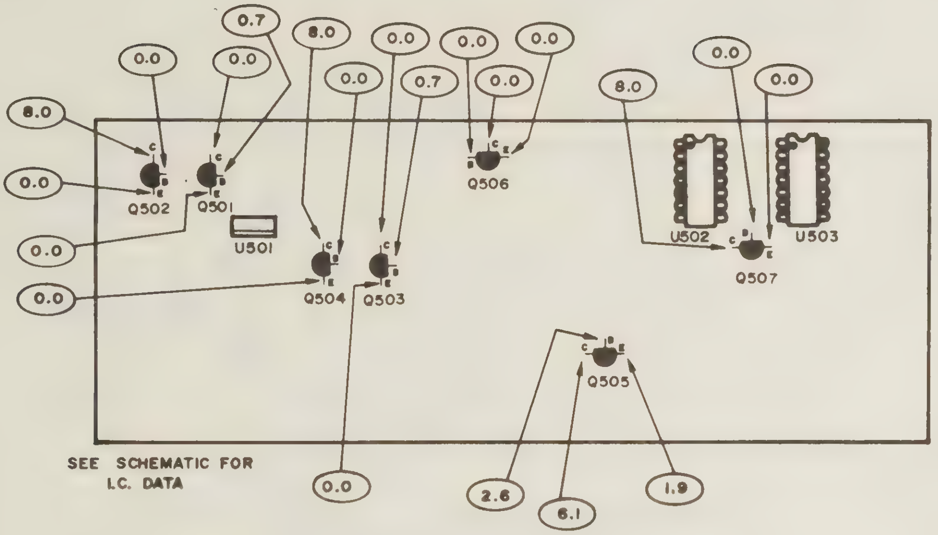
SEE SCHEMATIC FOR I.C. DATA

DISTRIBUTION BOARD LEVELS

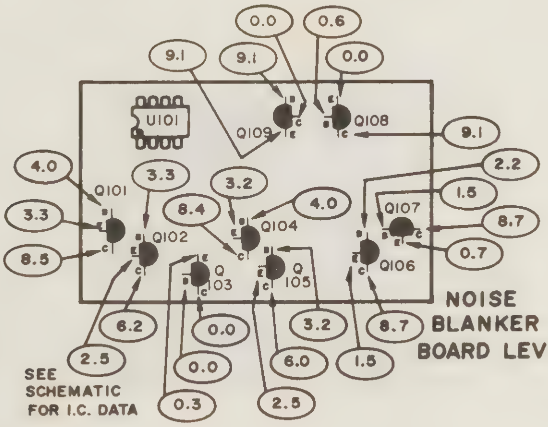


SEE SCHEMATIC FOR I.C. DATA

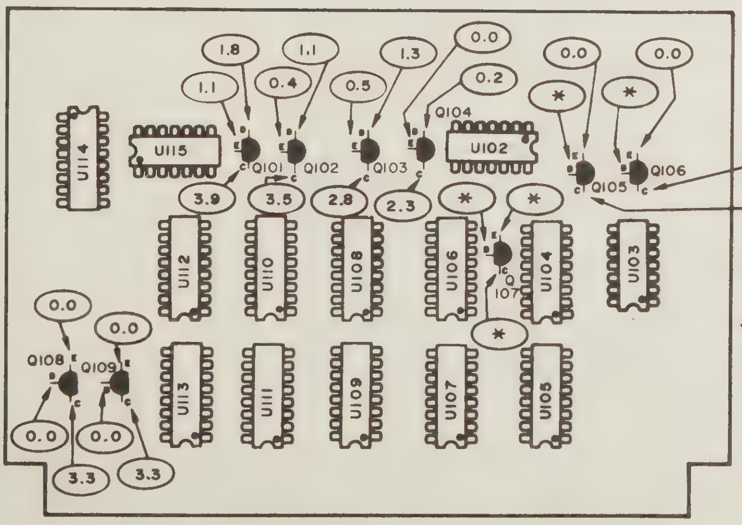
AUDIO BOARD LEVELS



PTO CONTROL BOARD LEVELS



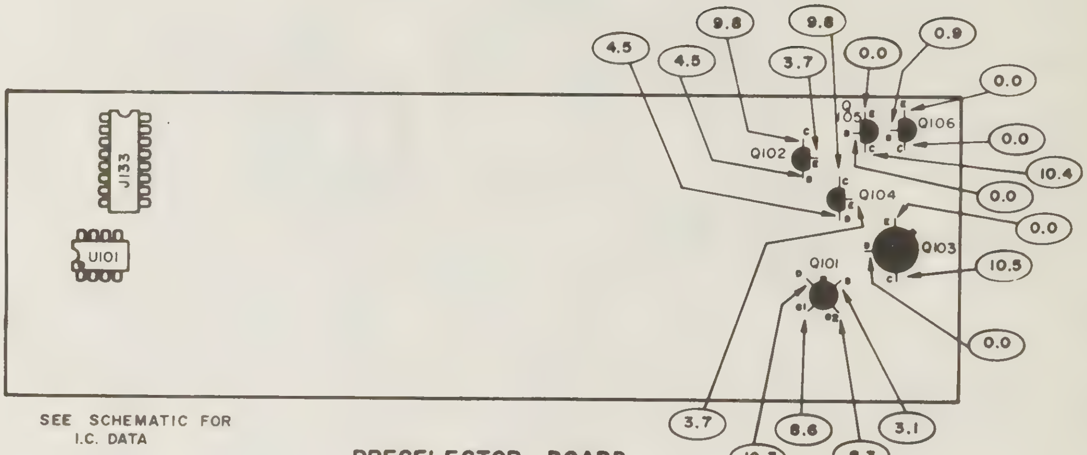
NOISE BLANKER BOARD LEVELS



SEE SCHEMATIC FOR I.C. DATA

*PRESENCE OF SQUARE WAVE MAKES READING MEANINGLESS

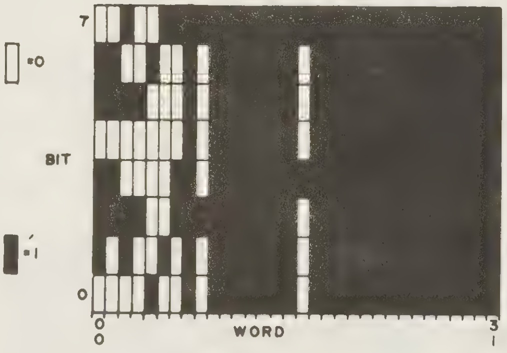
COUNTER BOARD LEVELS



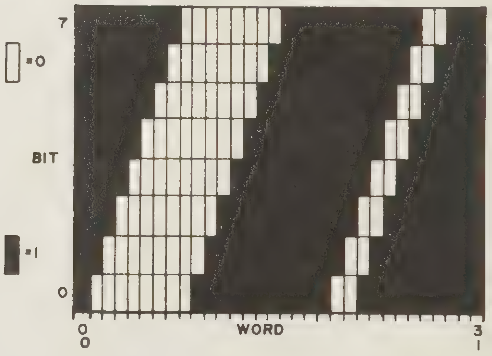
SEE SCHEMATIC FOR
I.C. DATA

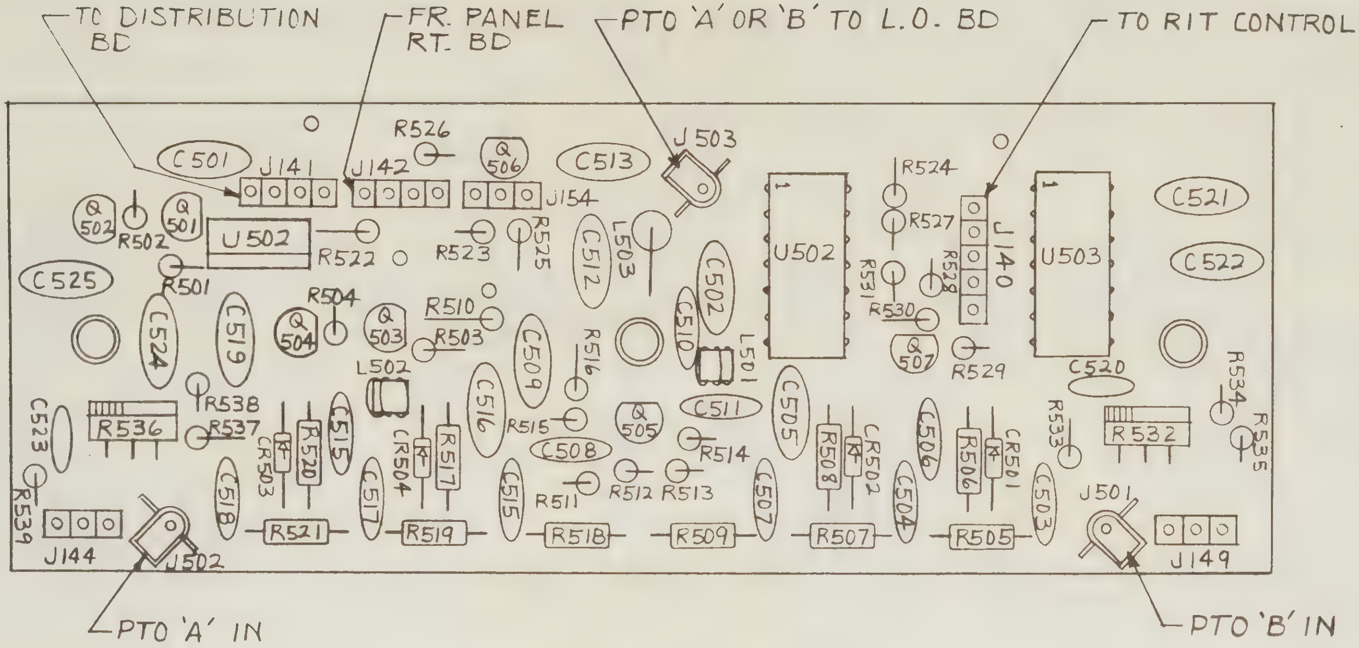
PRESELECTOR BOARD
LEVELS

U105 MEMORY MAP



U107 MEMORY MAP





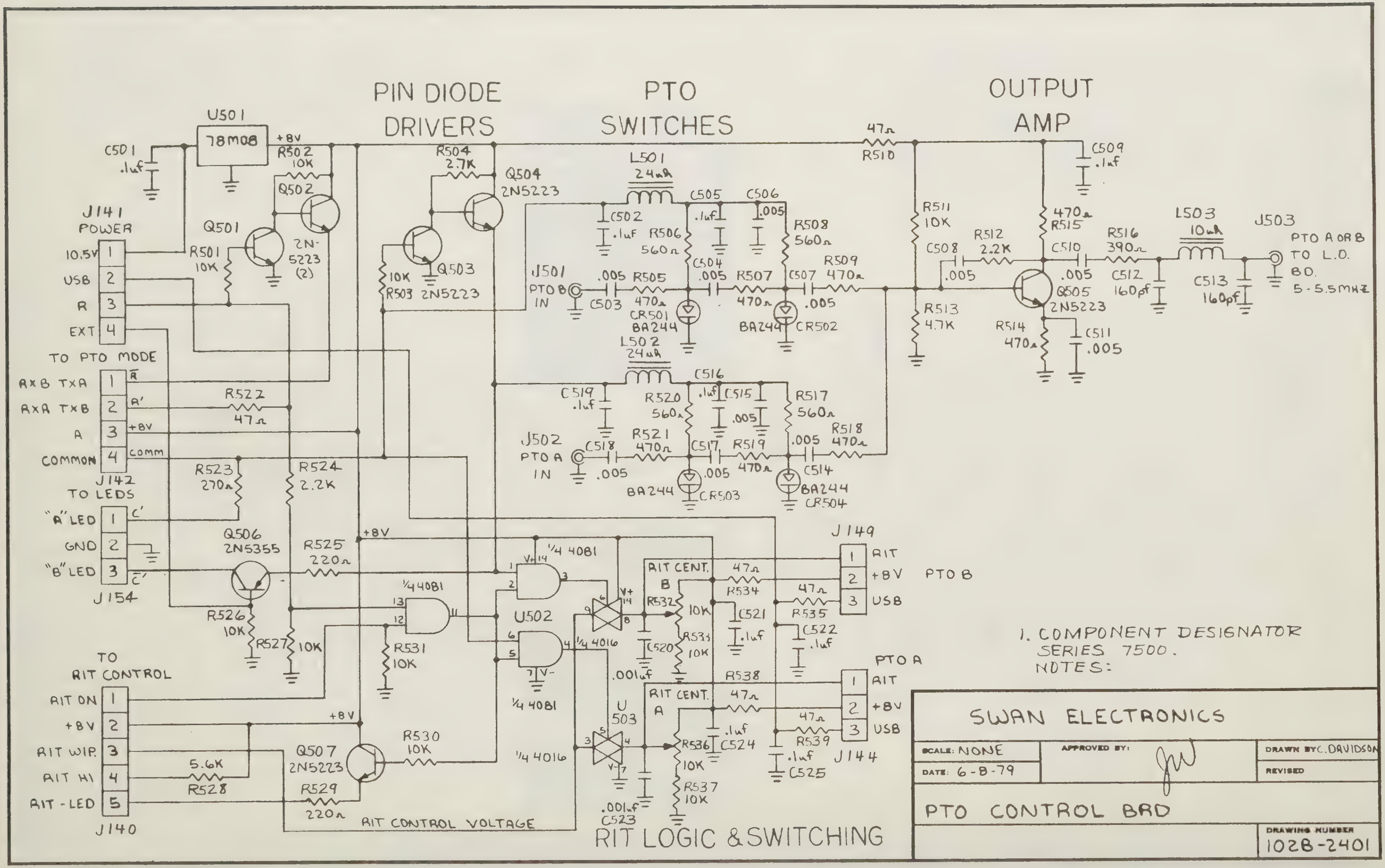
2. COMPONENT DESIGNATORS: 7500 SERIES
1. REF. ELEC. ASSY 102BX-2001
REF. CIRCUIT BD 102BX-2101
REF. DRILL DWG 102BX-2201
NOTES:

SWAN ELECTRONICS

SCALE: ~	APPROVED BY: JW	DRAWN BY AJG
DATE: 9-18-79		REVISED

PTO CONTROL BOARD LAYOUT

DRAWING NUMBER 102BX-2301



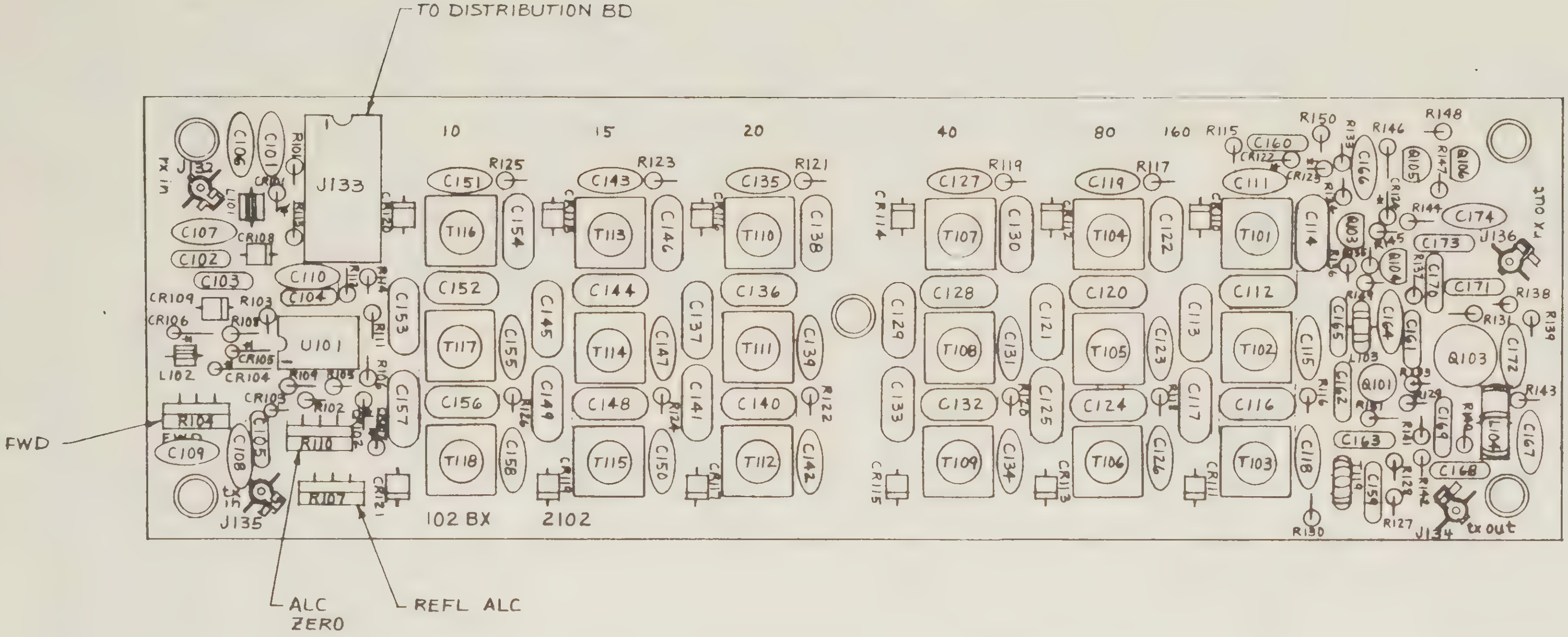
NOTES:
1. COMPONENT DESIGNATOR SERIES 7500.

SWAN ELECTRONICS

SCALE: NONE
DATE: 6-8-79
APPROVED BY: JW
DRAWN BY: C. DAVIDSON
REVISED

PTO CONTROL BRD

DRAWING NUMBER 102B-2401



2. COMPONENT DESIGNATION SERIES 8000 .
1. REF. ELECT. ASSY 102 BX- 2002
REF. CIRCUIT BD 102 BX- 2102
REF. DRILL DWG 102 BX- 2202
NOTES:

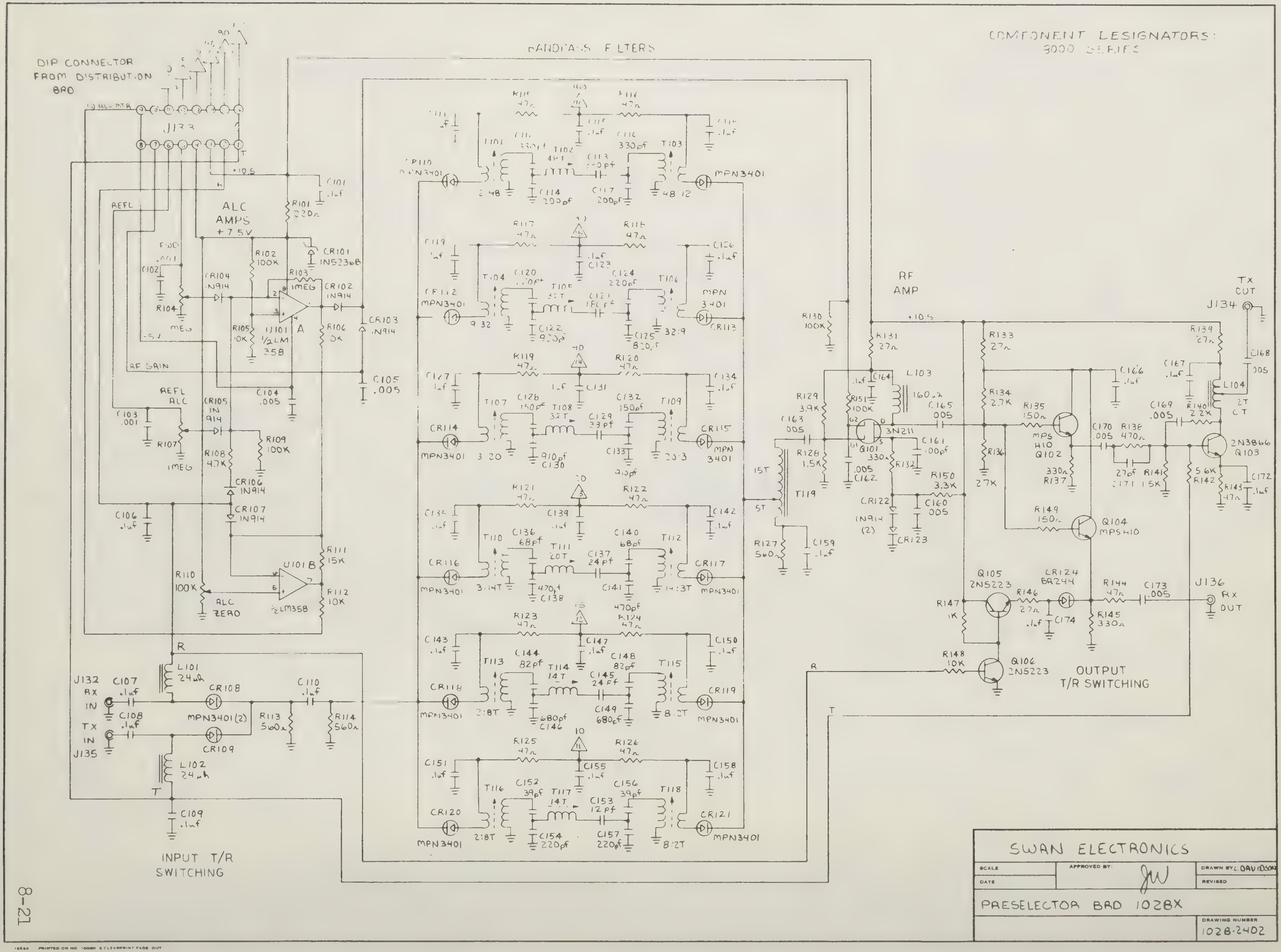
SWAN ELECTRONICS

SCALE NONE
DATE: 7-24-79
APPROVED BY. JW
DRAWN BY AJG
REVISED

PRESELECTOR BOARD
LAYOUT

REV
DRAWING NUMBER 102 BX-2302

COMPONENT DESIGNATORS:
3000 SERIES



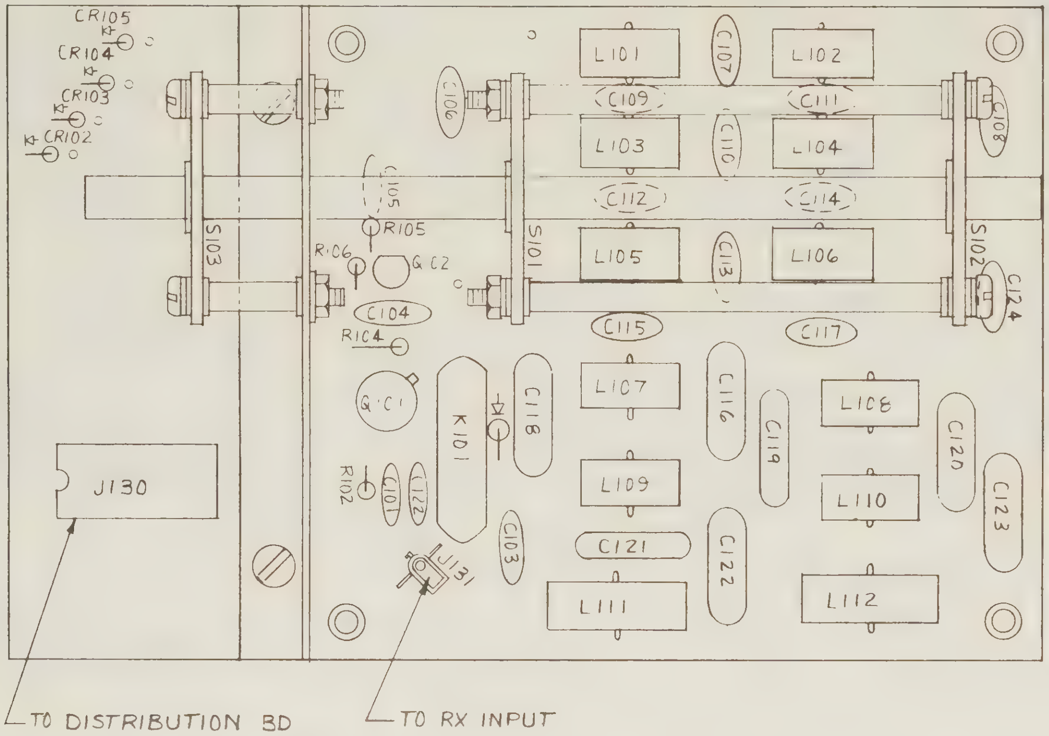
SWAN ELECTRONICS

SCALE	APPROVED BY: JW	DRAWN BY L. DAVIDSON
DATE		REVISED

PRESELECTOR BRD 1028X

DRAWING NUMBER
1028-2402

18884 PRINTED ON NO 1000H CLEARPRINT FADE OUT



2. COMPONENT DESIGNATORS 3000 SERIES.
1. REF. ELEC. ASSY 102BX-2003
REF. CIRCUIT BD 102BX-2103
REF. DRILL DWG 102BX-2203
NOTES:

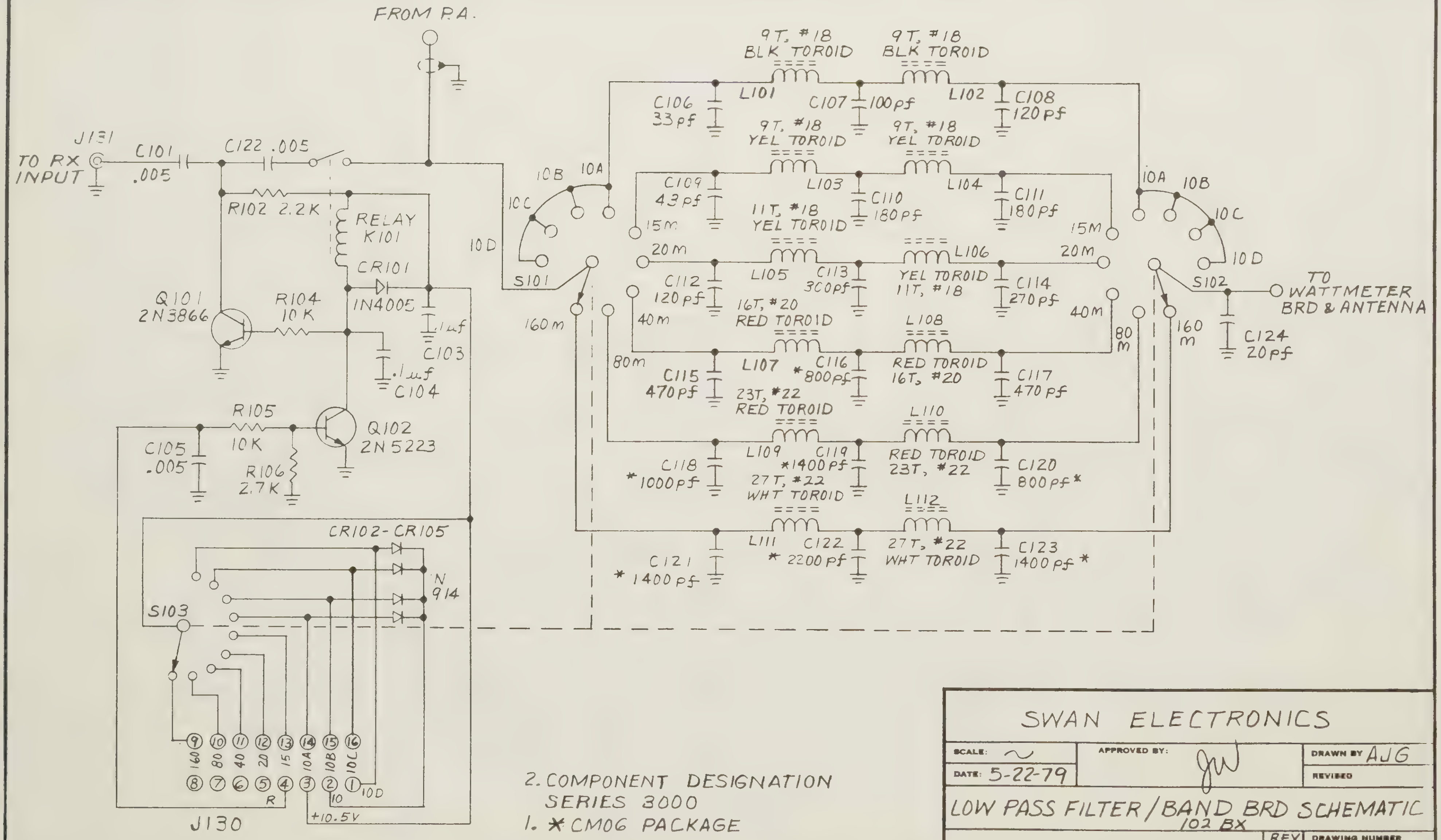
SWAN ELECTRONICS

SCALE: ~	APPROVED BY: JW	DRAWN BY AJG
DATE: 9-18-79		REVISED

B'SWITCH & L.P. FILT. BD LAYOUT

DRAWING NUMBER 102BX-2303

17282 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATION
SERIES 3000
1. ✱ CM06 PACKAGE
NOTES:

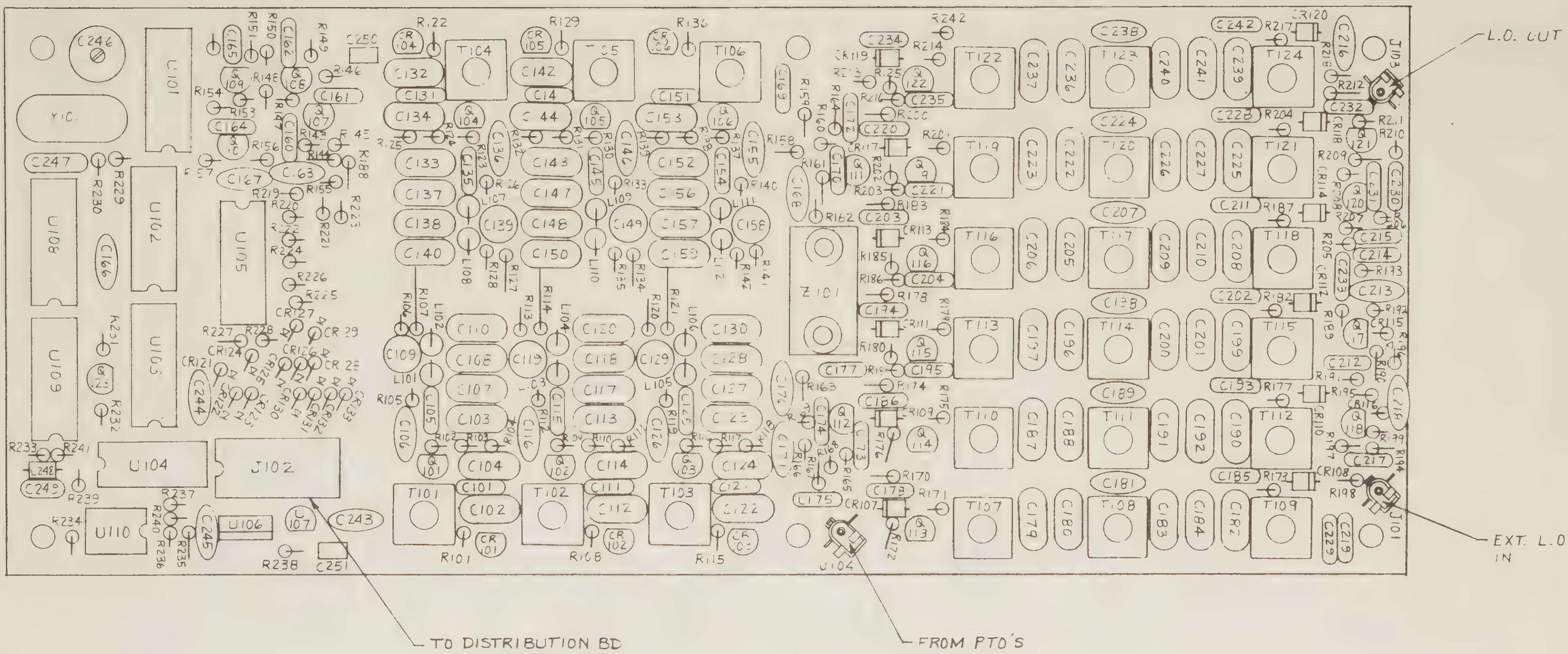
SWAN ELECTRONICS

SCALE: ~
DATE: 5-22-79
APPROVED BY: JW
DRAWN BY AJG
REVISED

LOW PASS FILTER / BAND BRD SCHEMATIC
102 BX

REV
DRAWING NUMBER 102BX-2403

11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS: 5000 SERIES

1. REF. ELECT. ASSY 102BX-2004
REF. CIRCUIT BOARD 102BX-2104
REF. DRILL DWG 102BX-2204

NOTES.

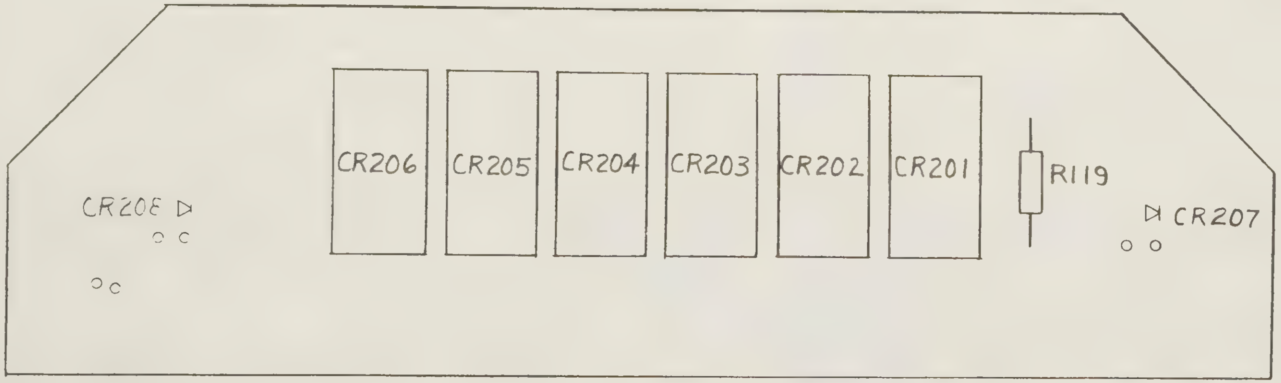
SWAN ELECTRONICS

SCALE NONE	APPROVED BY jw	DRAWN BY AJG
DATE: 7-30-79		REVISED

L.O. BOARD COMPONENT LAYOUT

REV. DRAWING NUMBER 102BX-2304

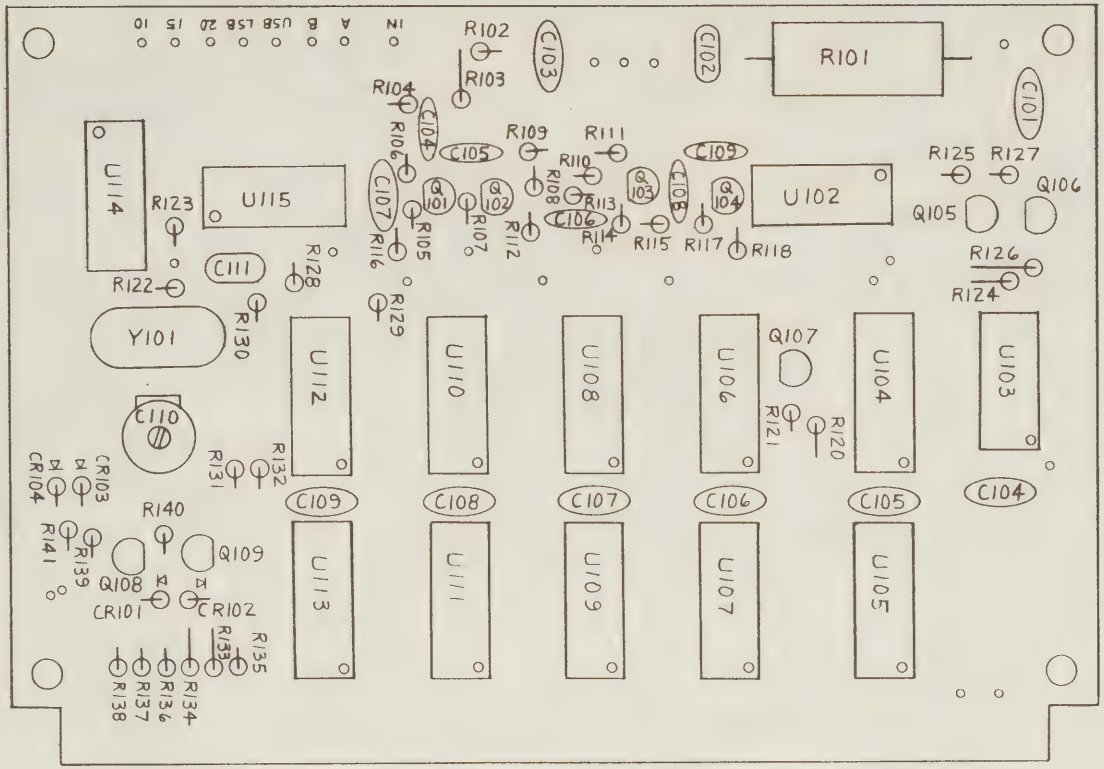
PRINTED ON NO. 1000M CLEARPRINT FADE OUT



2. COMPONENT DESIGNATOR SERIES 7200
1. REF. ELECT. ASSY 102BX-2006
REF. CIRCUIT BD 102BX-2106
REF. DRILL DWG 102BX-2206
NOTES:

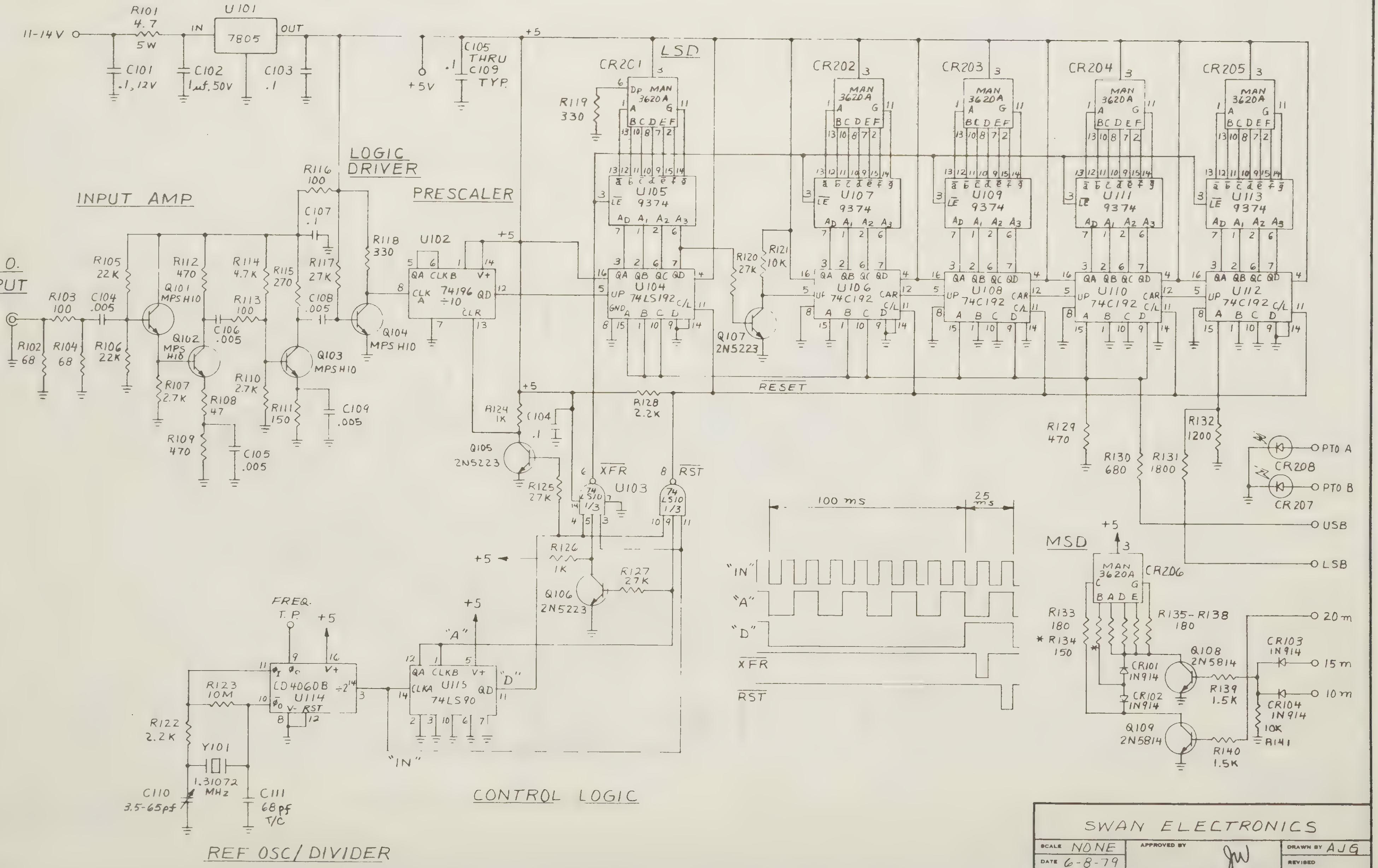
SWAN ELECTRONICS
SCALE: ~
DATE: 9-19-79
APPROVED BY: JW
DRAWN BY AJG
REVISED
DISPLAY BOARD LAYOUT
DRAWING NUMBER 102BX-2306

11X17 PRINTED ON NO. 1000H CLEARPRINT



2. COMPONENT DESIGNATORS 7100 SERIES
1. REF. ELEC. ASSY 102BX-2005
REF. CIRCUIT BD 102BX 2105
REF. DRILL DWG 102BX 2205
NOTES:

SWAN ELECTRONICS
SCALE: ~
DATE: 9-19-79
APPROVED BY: JW
DRAWN BY AJG
REVISED
COUNTER BOARD LAYOUT
DRAWING NUMBER 102BX-2305



NOTES.
1. COUNTER REF. DESIGNATOR SERIES 7100
2. DISPLAY REF. DESIGNATOR SERIES 7200

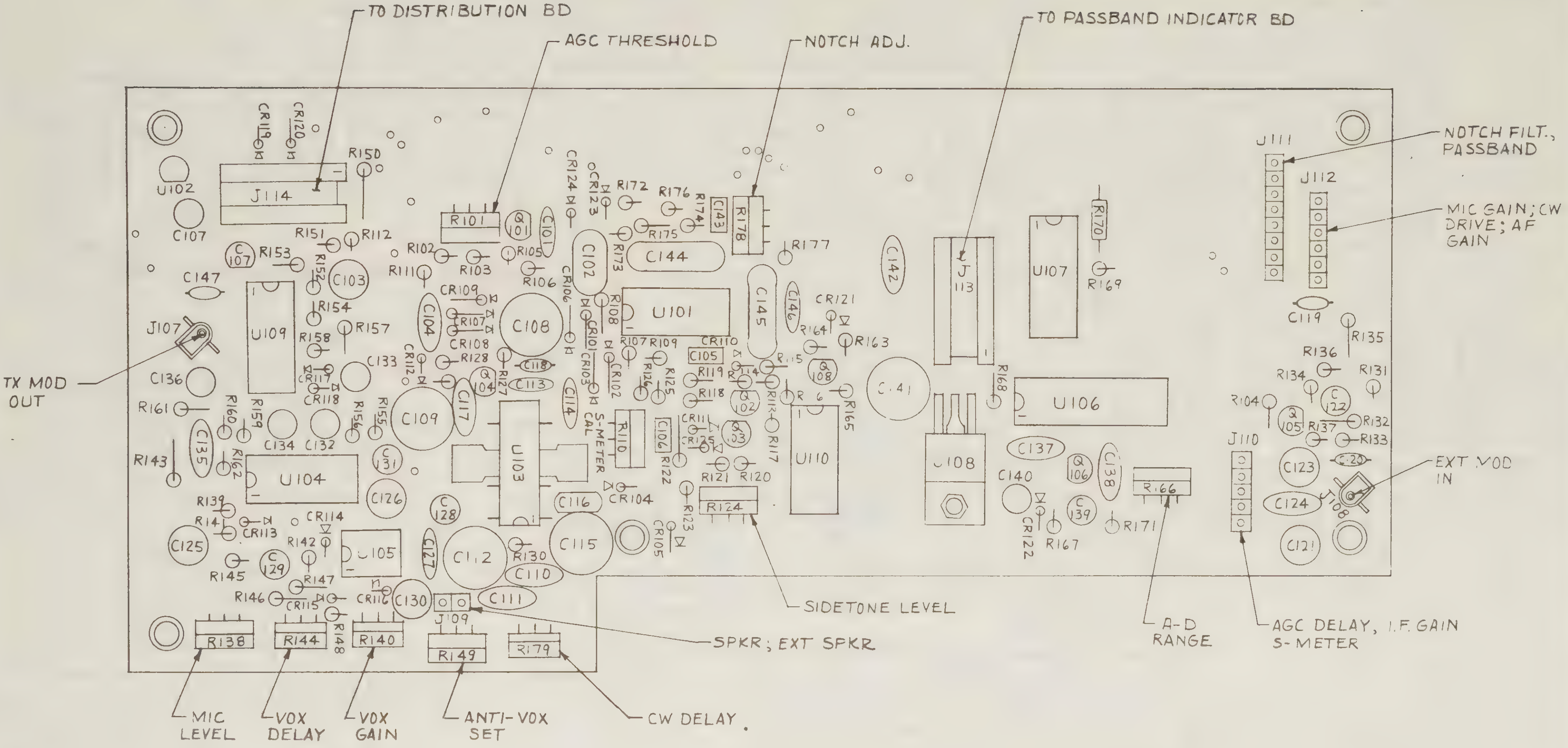
SWAN ELECTRONICS

SCALE NONE
DATE 6-8-79
APPROVED BY JW
DRAWN BY AJG
REVISED

COUNTER / DISPLAY BD SCHEMATIC 102 BX

REV
DRAWING NUMBER 102BX-2405

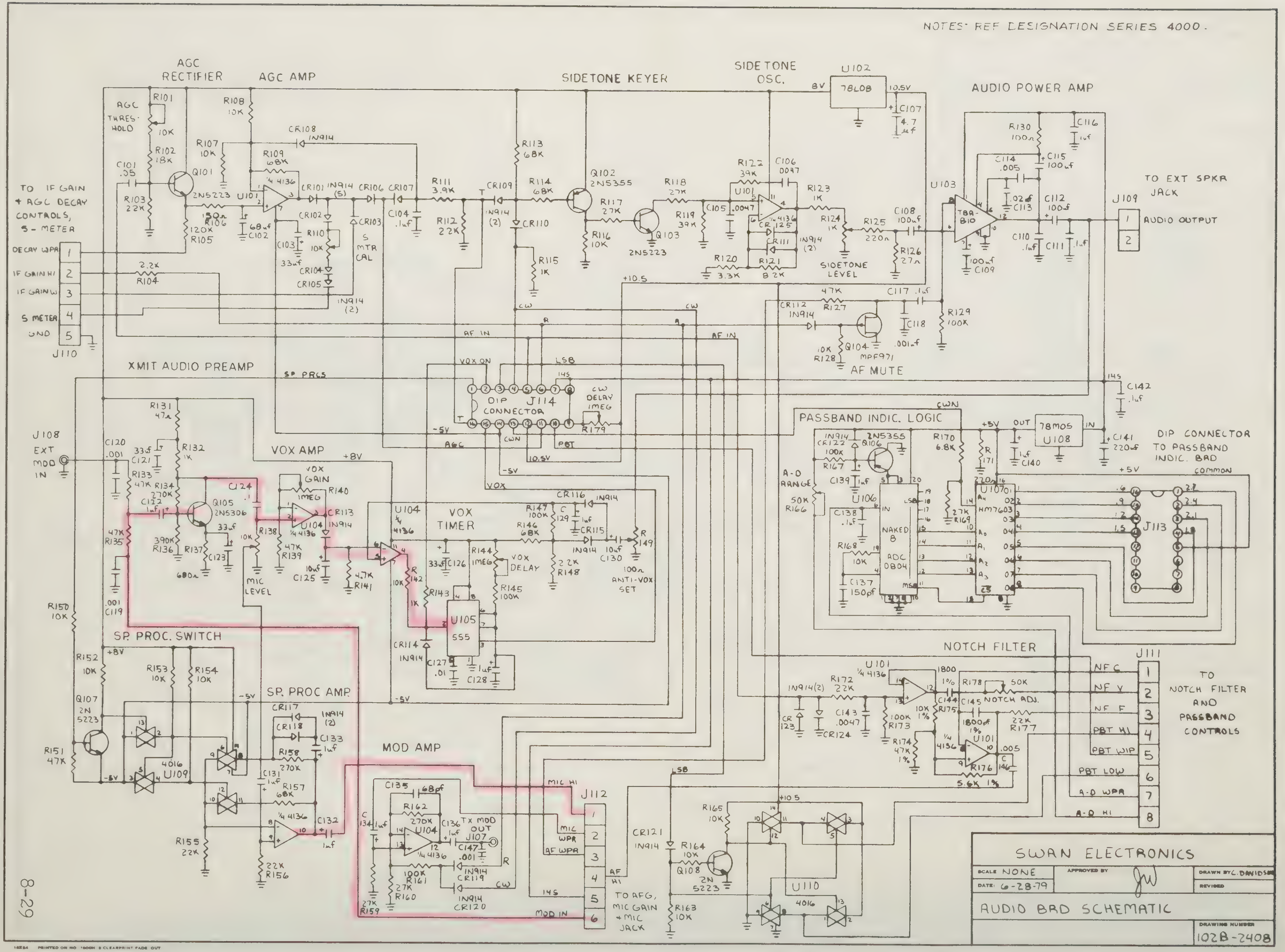
18284 PRINTED ON NO 1000H 8 CLEARPRINT FADE OUT

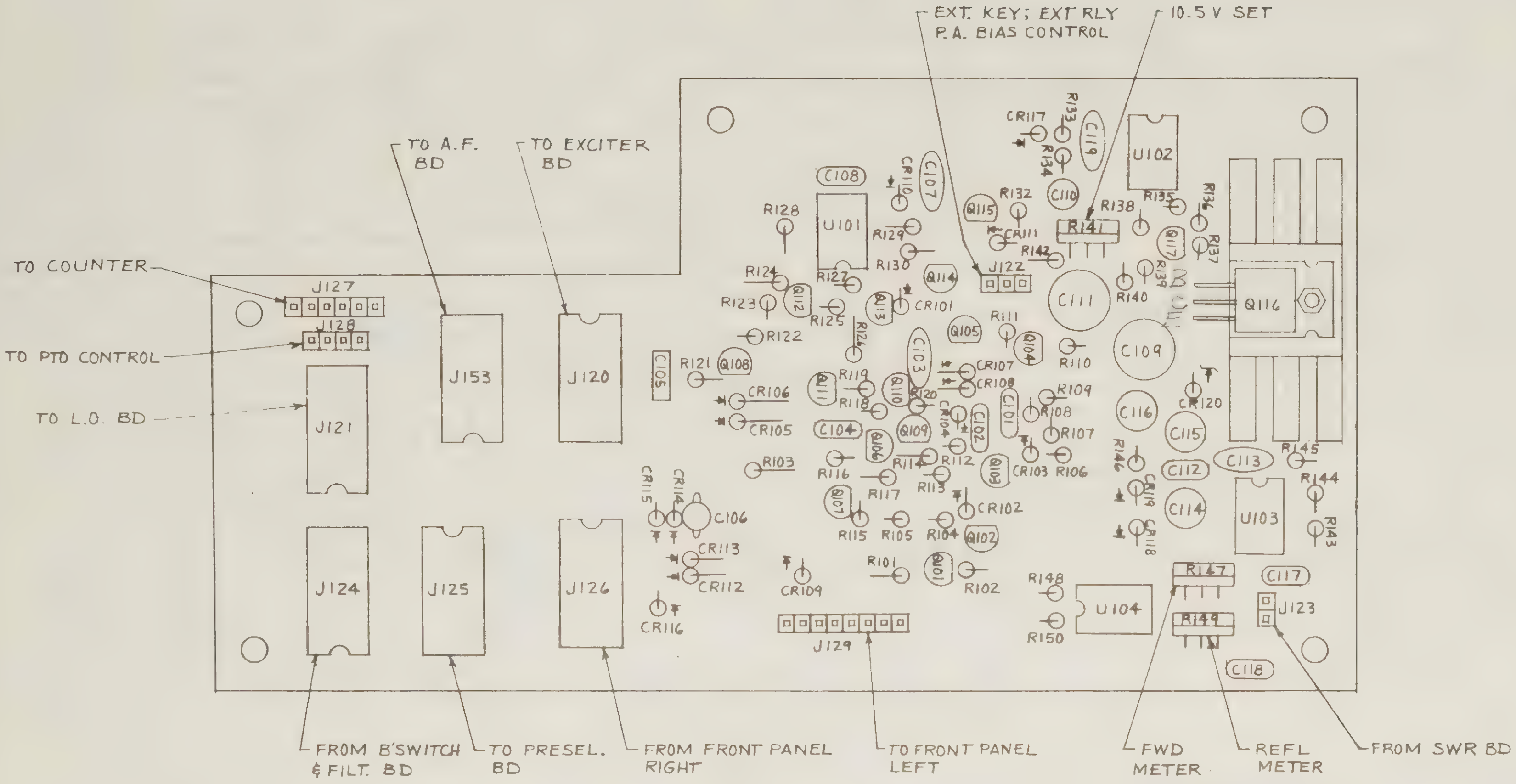


2. COMPONENT DESIGNATORS 4000 SERIES.
1. REF. ELECT. ASSY 102BX-2008
REF. CIRCUIT BOARD 102BX-2108
REF. DRILL DWG 102BX-2208
NOTES:

SWAN ELECTRONICS		
SCALE: NONE DATE:	APPROVED BY: JW	DRAWN BY AJG REVISED
AUDIO BOARD LAYOUT		
		DRAWING NUMBER 102BX-2308

17X22 PRINTED ON NO 1000H CLEARPRINT


NOTES: REF DESIGNATION SERIES 4000.
AGC RECTIFIER
AGC AMP
SIDETONE KEYER
SIDETONE OSC.
AUDIO POWER AMP
AGC THRES-HOLD
S MTR CAL
SIDETONE LEVEL
AF MUTE
TO EXT SPKR JACK
AUDIO OUTPUT
J109
TO IF GAIN + AGC DECAY CONTROLS, S-METER
DECAY WPR
IF GAIN HI
IF GAIN W
S METER
GND
J110
XMIT AUDIO PREAMP
DIP CONNECTOR
J114
CW DELAY 1MEG
PASSBAND INDIC. LOGIC
DIP CONNECTOR TO PASSBAND INDIC. BRD
COMMON
J113
A-D RANGE
NAKED 8 ADC 0804
HM7603
78M05 U108
78L08 U102
TBA-810 U103
J108 EXT MOD IN
VOX AMP
VOX GAIN 1MEG
MIC LEVEL
VOX TIMER
VOX DELAY
ANTI-VOX SET
U105 555
SP. PROC. SWITCH
SP. PROC AMP
MOD AMP
TX MOD OUT J107
J112
MIC HI
MIC WPR
AF WPR
AF HI
MOD IN
TO AFG, MIC GAIN + MIC JACK
NOTCH FILTER
NOTCH ADJ.
J111
NF C
NF Y
NF F
PBT HI
PBT WIP
PBT LOW
A-D WPR
A-D HI
TO NOTCH FILTER AND PASSBAND CONTROLS
U110 4016
U109 4016
SWAN ELECTRONICS
SCALE NONE
DATE: 6-28-79
APPROVED BY JW
DRAWN BY C. DAVIDSON
REVISED
AUDIO BRD SCHEMATIC
DRAWING NUMBER 102B-2408



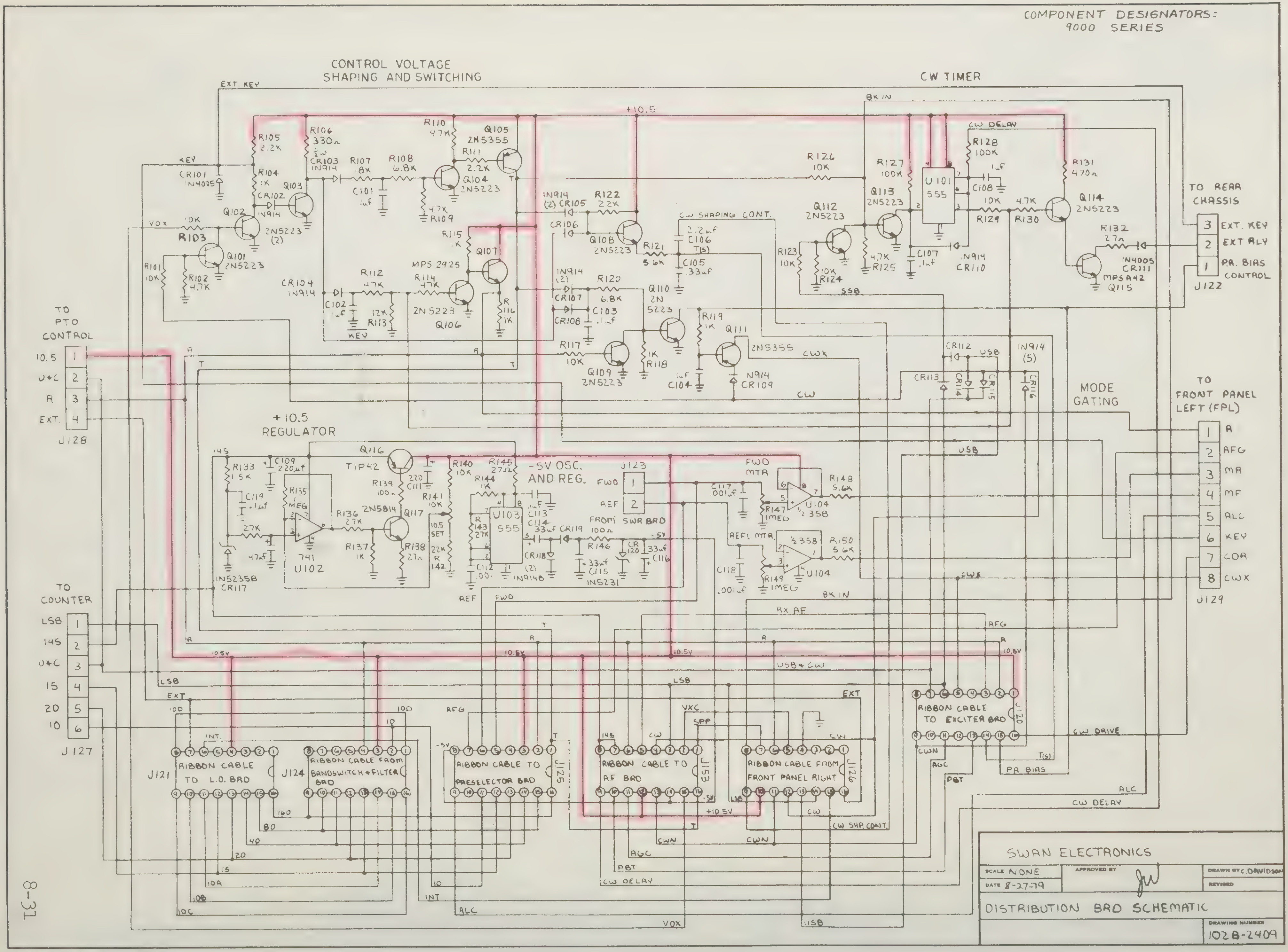
2. COMPONENT DESIGNATORS: 9000 SERIES

1. REF. ELECT. ASSY 102BX-2009
REF. CIRCUIT BD 102BX-2109
REF. DRILL DWG 102BX-2209
NOTES:

SWAN ELECTRONICS
SCALE NONE
DATE 7-25-79
APPROVED BY JW
DRAWN BY AJG
REVISED
DISTRIBUTION BOARD
COMPONENT LAYOUT
REV
DRAWING NUMBER 102BX-2309

PRINTED ON NO 1000H CLEARPRINT FADE OUT

COMPONENT DESIGNATORS:
9000 SERIES



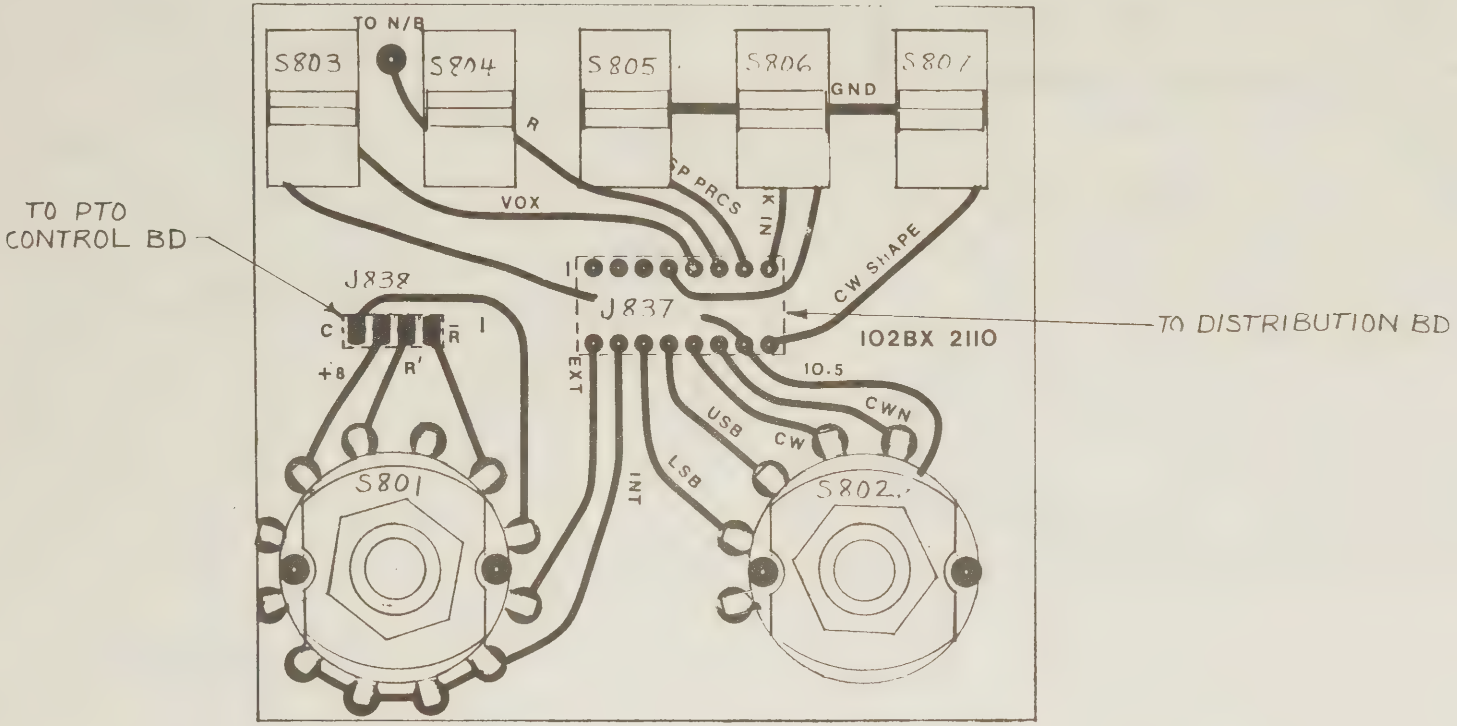
SWAN ELECTRONICS

SCALE NONE
DATE 8-27-79
APPROVED BY JW
DRAWN BY C. DAVIDSON
REVISED

DISTRIBUTION BRD SCHEMATIC

DRAWING NUMBER 102B-2409

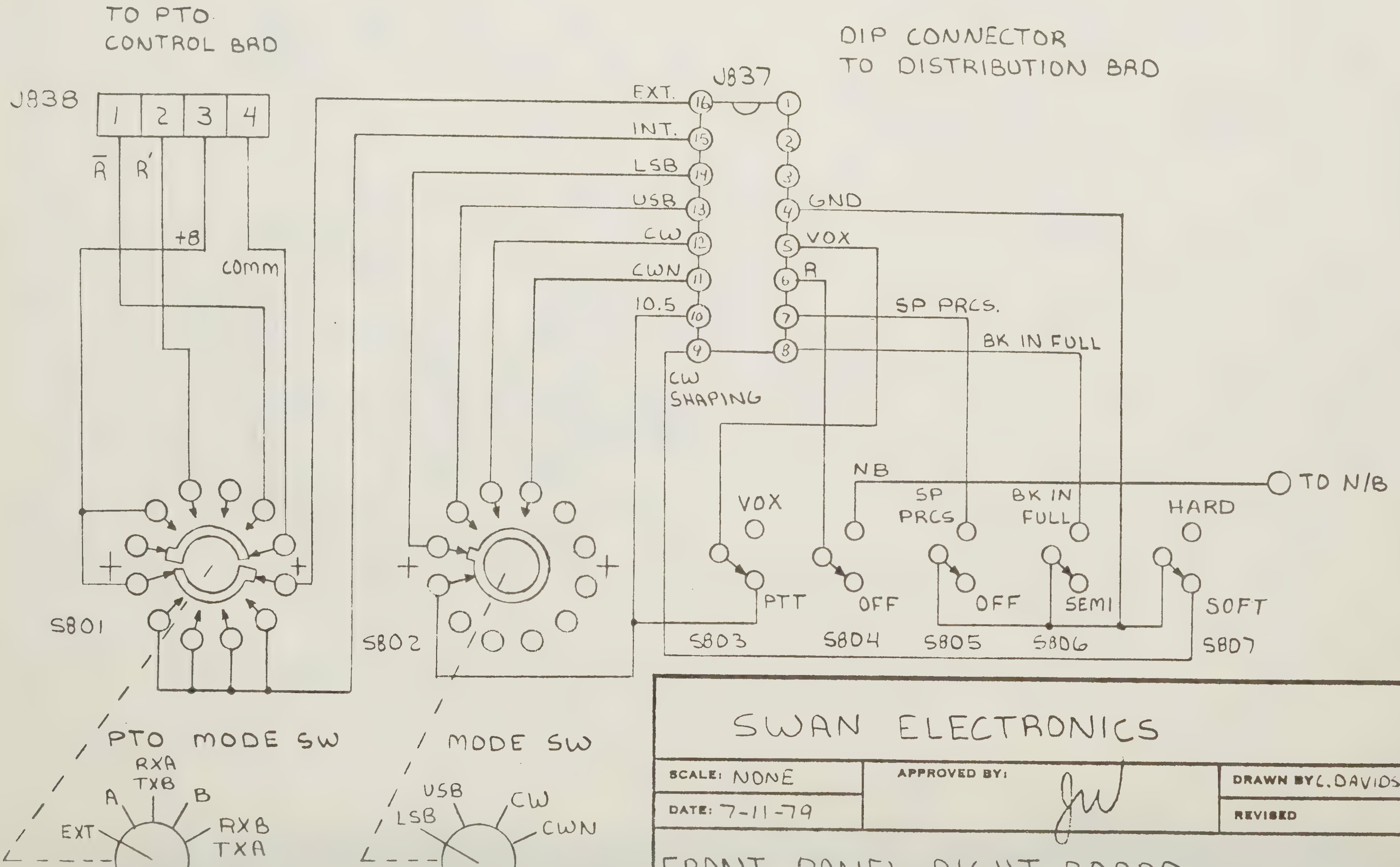
18X24 PRINTED ON NO 1000H 8 CLEARPRINT FADE OUT



3. J837 & J838 ON FARSIDE.
2. COMPONENT DESIGNATOR SERIES 7800.
1. REF. ELECTRICAL ASSY 102BX-2010
REF. CIRCUIT BOARD 102BX-2110
REF. DRILL DRAWING 102BX-2210
NOTES:

SWAN ELECTRONICS
SCALE: 2:1
DATE: 9-12-79
APPROVED BY: JW
DRAWN BY AJG
REVISED
FRONT PANEL RIGHT BD LAYOUT
DRAWING NUMBER 102BX-2310

11X17 PRINTED ON NO. 1000H CLEARPRINT



SWAN ELECTRONICS

SCALE: NONE

DATE: 7-11-79

APPROVED BY: JW

DRAWN BY C. DAVIDSON

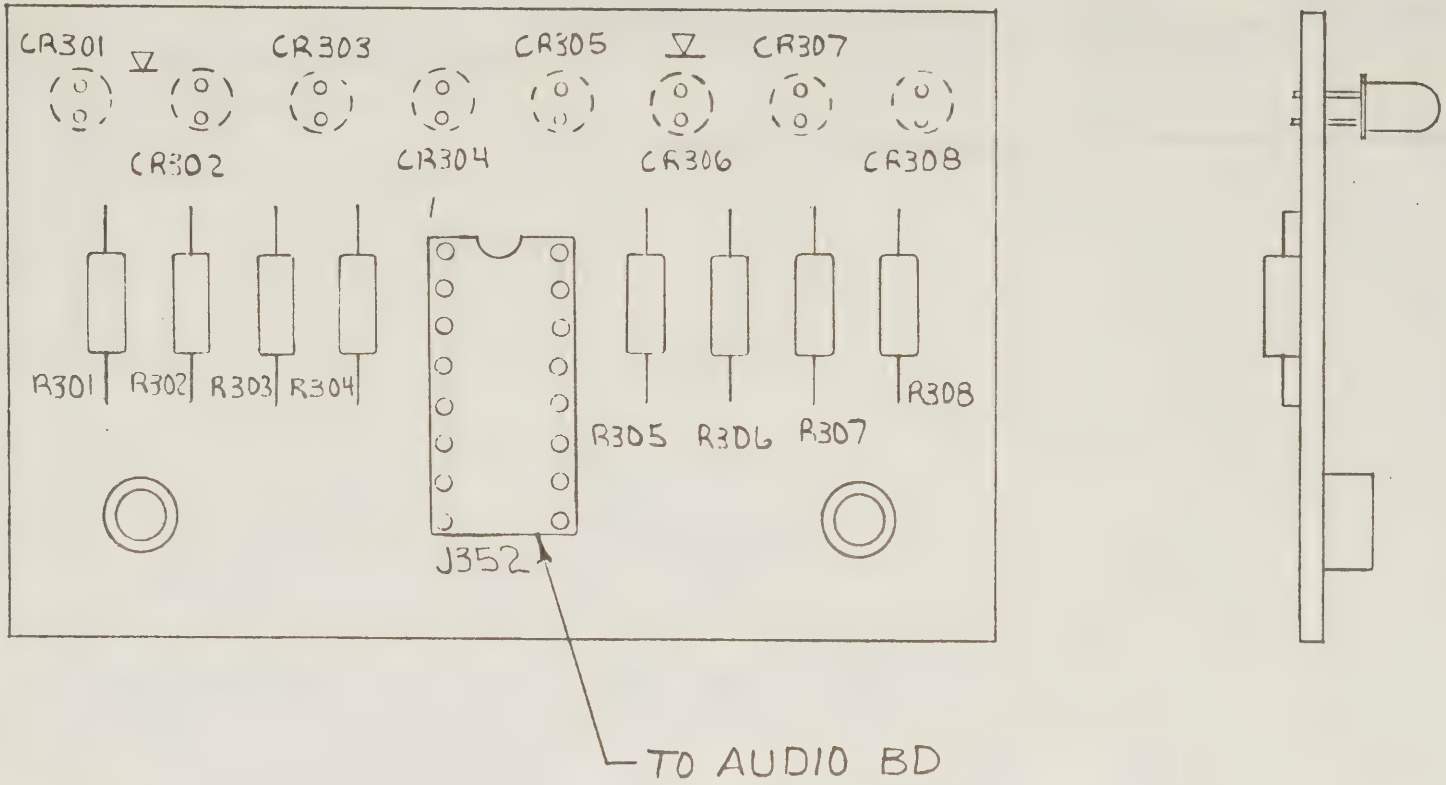
REVISED

FRONT PANEL RIGHT BOARD

COMPONENT DESIGNATORS (7800 SERIES)

DRAWING NUMBER 102BX-2410

8½x11 PRINTED ON NO. 1000H-10 CLEARPRINT FADE-OUT



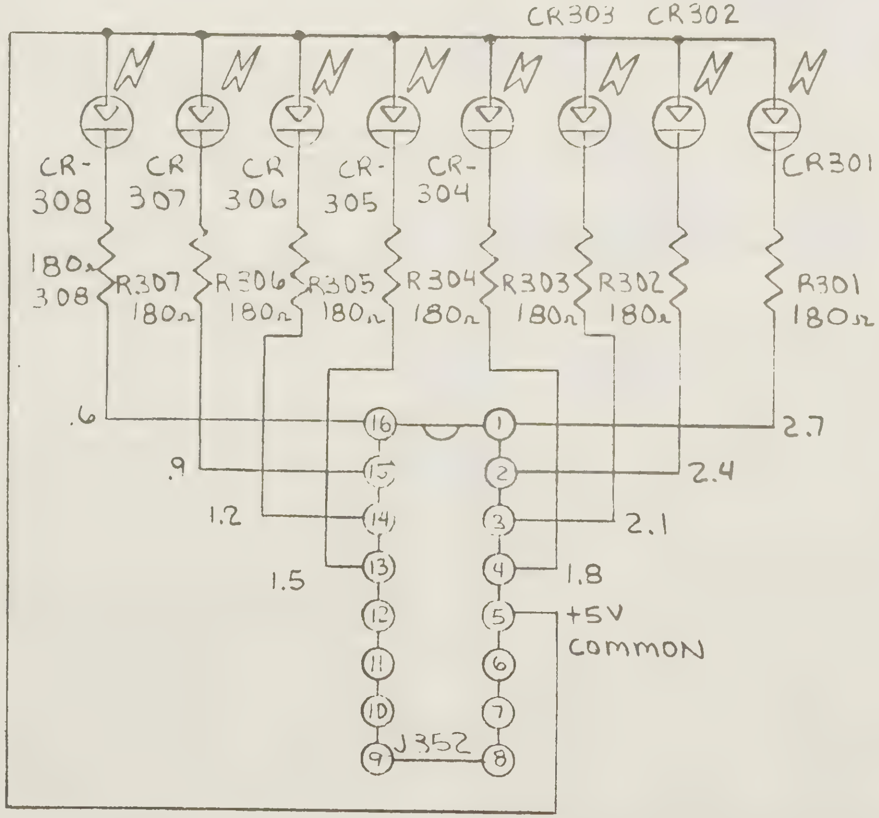
3. REF. ELECTRICAL ASSY 102BX-2011
REF. CIRCUIT BOARD 102BX-2111
REF. DRILL DWG 102BX-2211
2. LEDS MOUNTED FARSIDE.
1. COMPONENT DESIGNATORS 7300 SERIES
NOTES

SWAN ELECTRONICS

SCALE: NONE
APPROVED BY:
DRAWN BY AJG
DATE:
REVISED

PASSBAND IND. BD LAYOUT
102 BX

DRAWING NUMBER
102BX-2311



DIP CONNECTOR
FROM AUDIO BRD.

SWAN ELECTRONICS

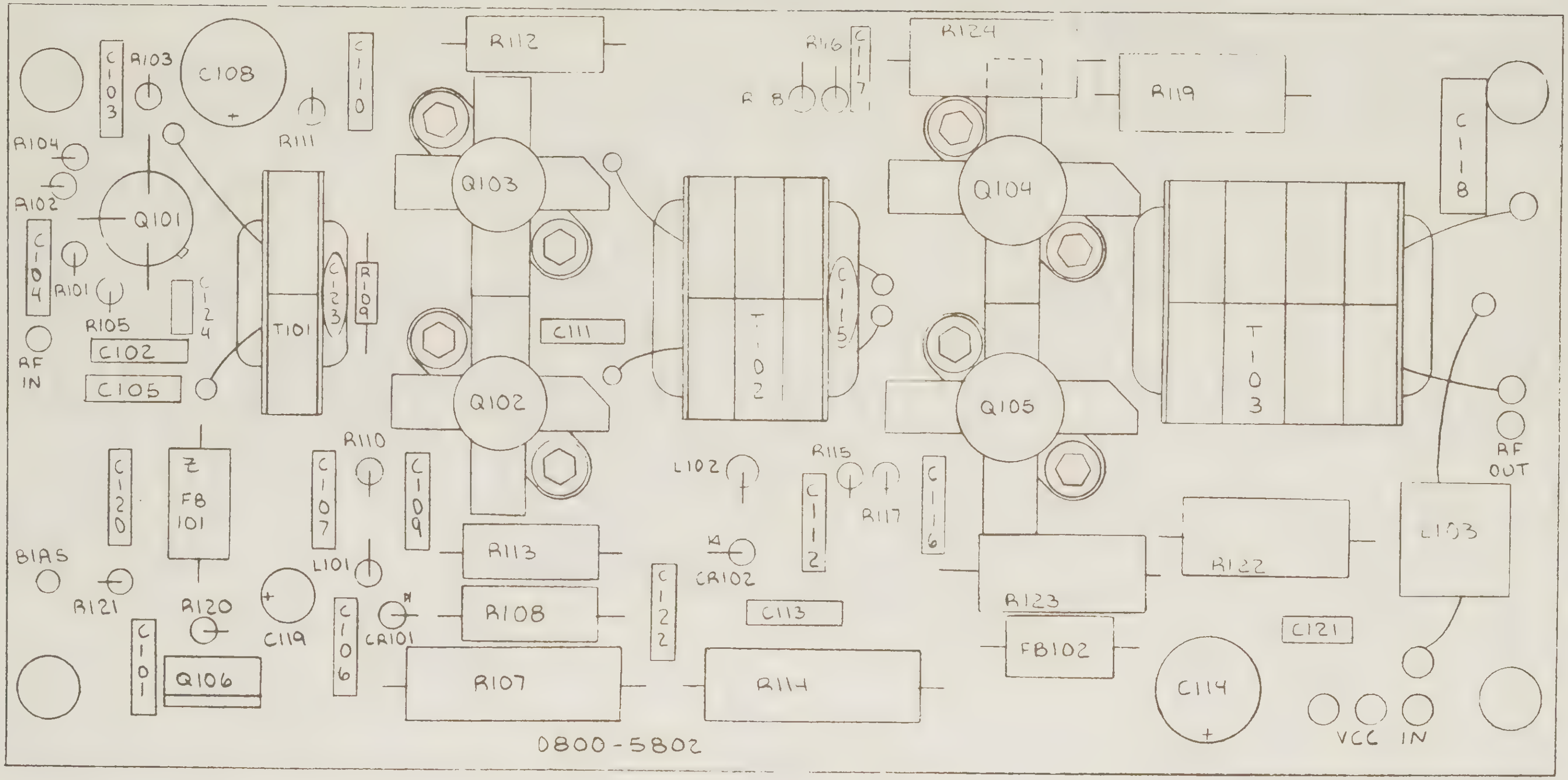
SCALE: NONE
DATE: 7-10-79
APPROVED BY: JW
DRAWN BY C. DAVIDSON
REVISED

PASSBAND INDICATOR BOARD SCHEMATIC

DRAWING NUMBER
102B-2411

1. COMPONENT DESIGNATORS 7300 SERIES
NOTES:

8½ X 11 PRINTED ON NO. 1000H CLEARPRINT



SWAN ELECTRONICS

SCALE: NONE

DATE: 1-2-78

APPROVED BY:

DRAWN BY: C DAVIDSON

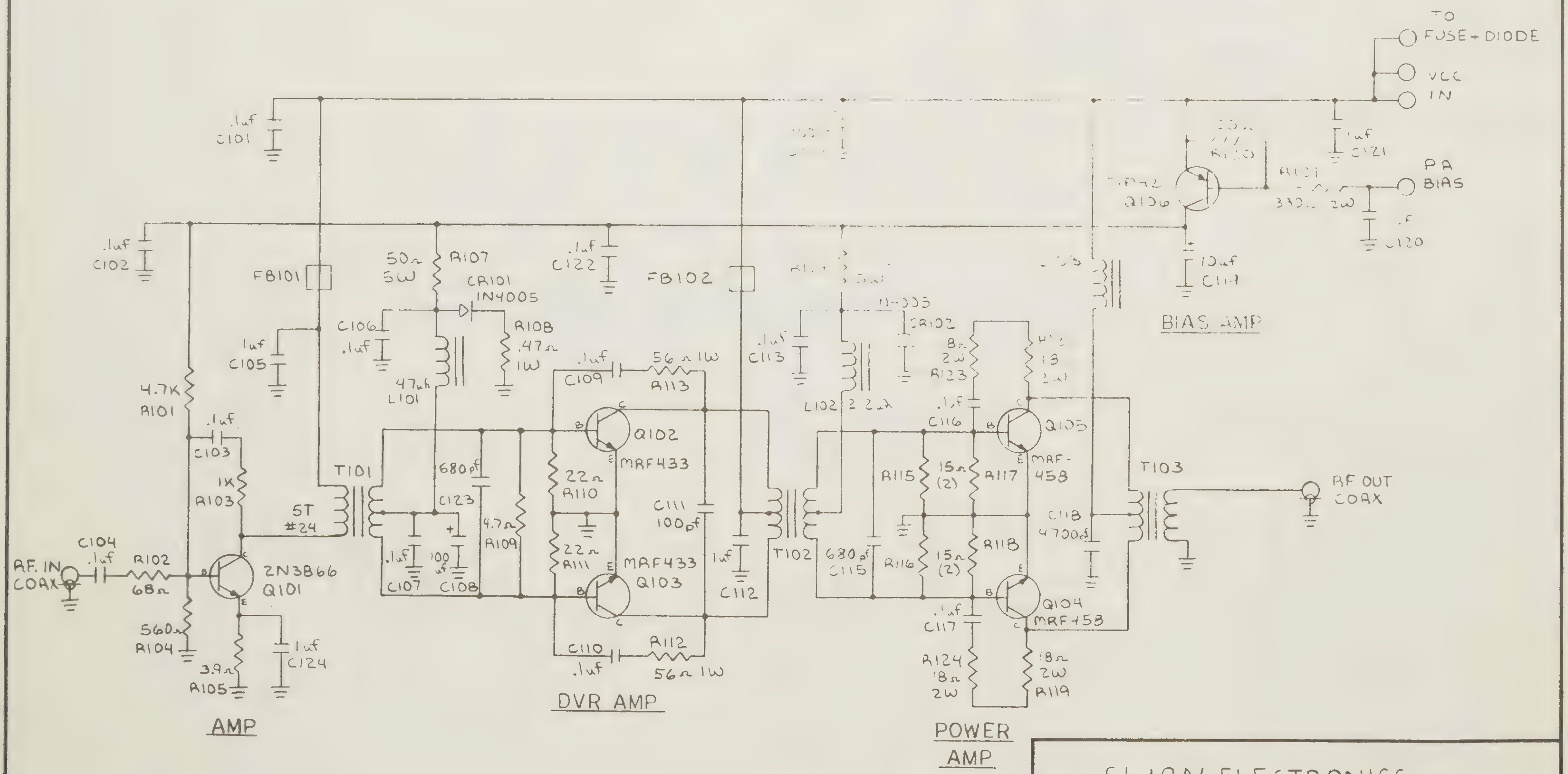
REVISED

POWER AMP BOARD LAYOUT

DRAWING NUMBER 0800-2302

NOTES:

1. COMPONENT DESIGNATORS (6000 SERIES)



SWAN ELECTRONICS

SCALE: NONE

DATE: 1-2-78

APPROVED BY:

DRAWN BY: C DAVIDSON

REVISED

POWER AMP BOARD SCHEMATIC

REV B

DRAWING NUMBER 0800-2402B

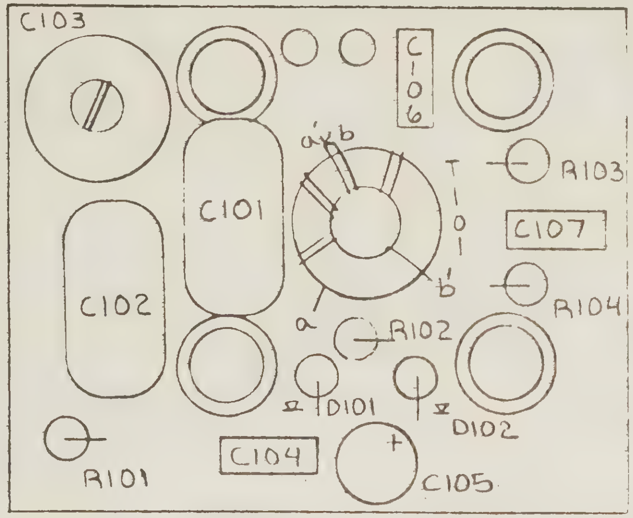
LTR	DESCRIPTION	DATE	APPD
ECN1 B	REPLACED ALL .05uf CAPS WITH .1uf CD 5-30-79	6-1-79	JW

REVISIONS

1. COMPONENT DESIGNATORS (6000 SERIES)

NOTES

11X17 PRINTED ON NO 1000H CLEARPRINT



SWAN ELECTRONICS

SCALE: NONE	APPROVED BY:	DRAWN BY C. DAVIDSON
DATE: 1-11-79		REVISED

SWR/WATTMETER BRD LAYOUT

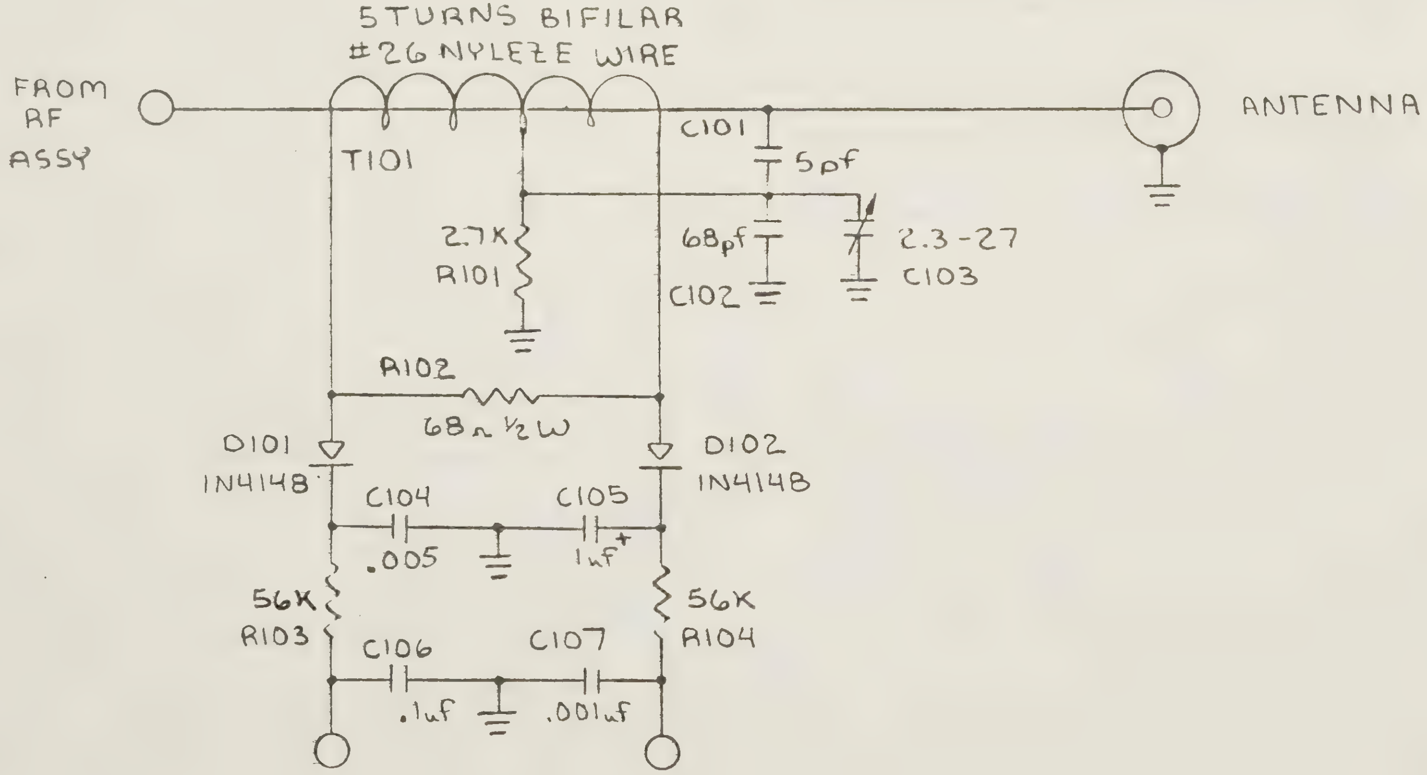
DRAWING NUMBER
0800-2303

1. COMPONENT DESIGNATORS (7700 SERIES)
NOTES:

8½X11 PRINTED ON NO. 1000H CLEARPRINT

REVISIONS

LTR	DESCRIPTION	DATE	APP'D
A	CHANGED C106 FROM .001 TO .1uF CD 8-24-79	8-24-79	Jw



SWAN ELECTRONICS

SCALE: NONE

APPROVED BY:

DRAWN BY: C DAVIDSON

DATE: 1-11-79

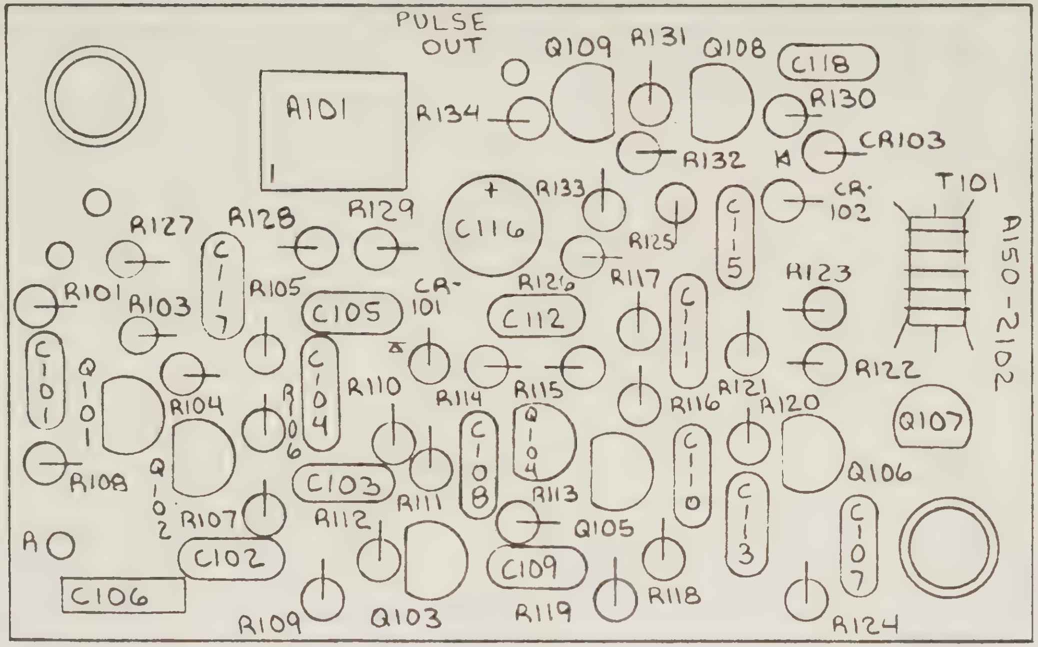
REVISED

SWR/WATTMETER BRD SCHEMATIC

DRAWING NUMBER
0800-2403A

1. COMPONENT DESIGNATORS (7700 SERIES)
NOTES

8½X11 PRINTED ON NO. 1000H CLEARPRINT



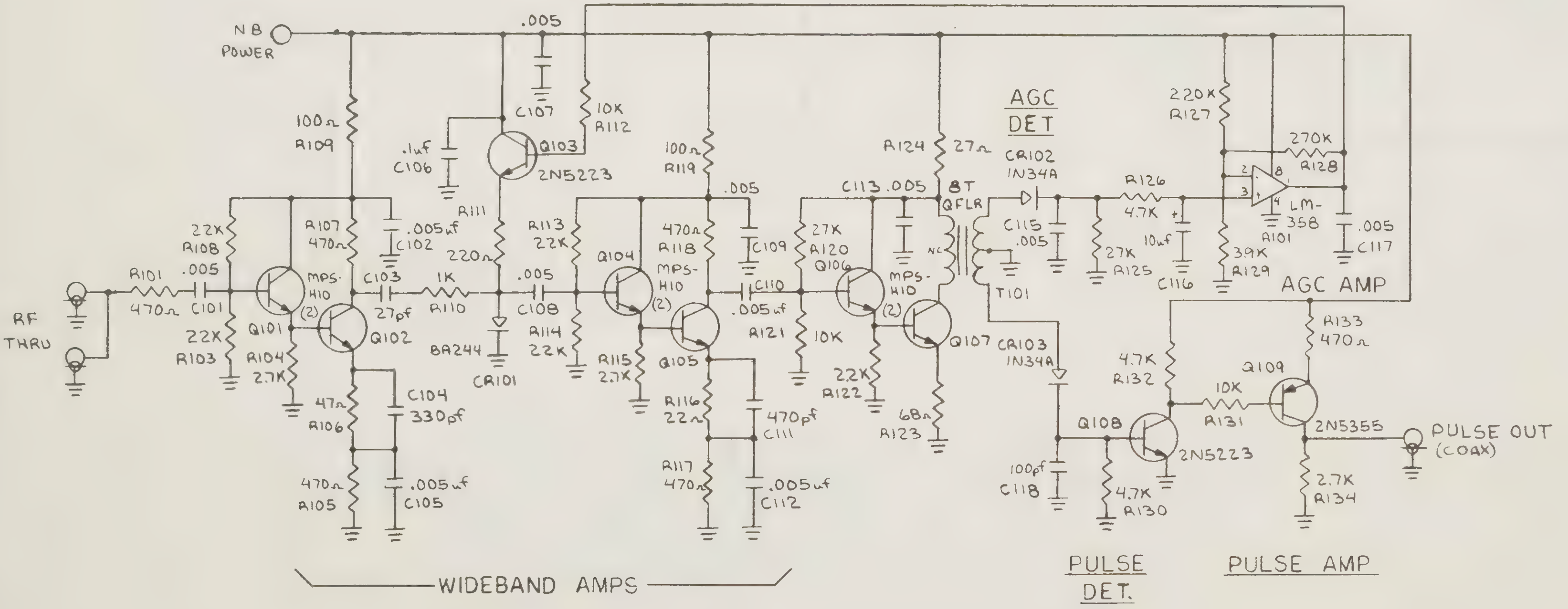
SWAN ELECTRONICS

SCALE: NONE	APPROVED BY:	DRAWN BY: C. DAVIDSON
DATE: 1-12-79		REVISED

PROJ.	A150 NOISE BLANKER BRD LAYOUT
0176	DRAWING NUMBER 0800-2304

1. COMPONENT DESIGNATORS (7400 SERIES)
NOTES

8½X11 PRINTED ON NO. 1000H CLEARPRINT



SWAN ELECTRONICS

SCALE: NONE
APPROVED BY: JW
DRAWN BY C.DAVID
DATE: 1-12-78
REVISED

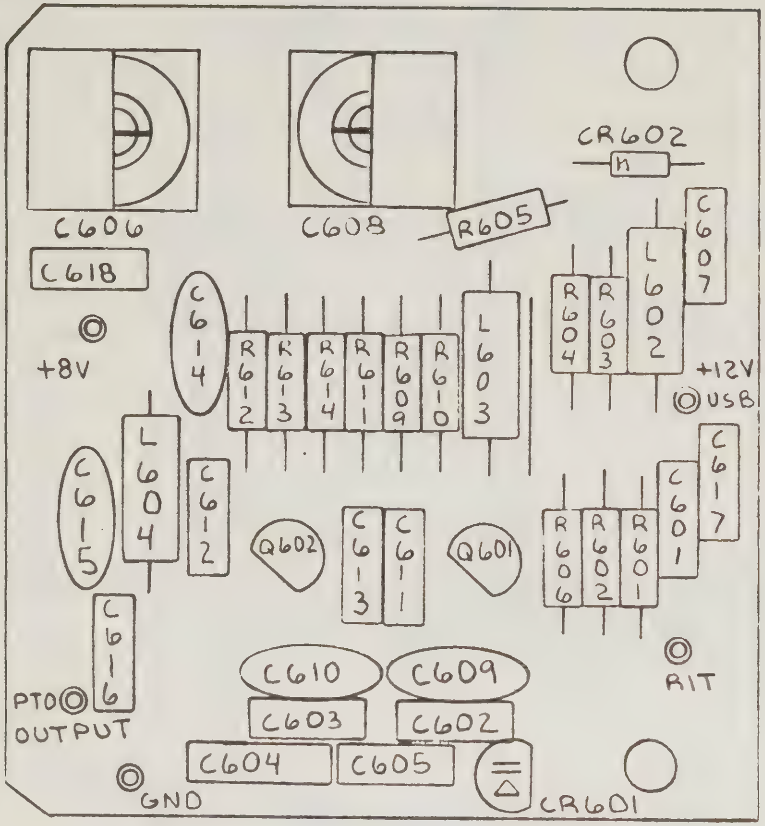
PROJ. 0176

A150 NOISE BLANKER BRD SCHEMATIC

DRAWING NUMBER 0800-2404A

1. COMPONENT DESIGNATORS (7400 SERIES)
NOTES

11X17 PRINTED ON NO. 1000H CLEARPRINT



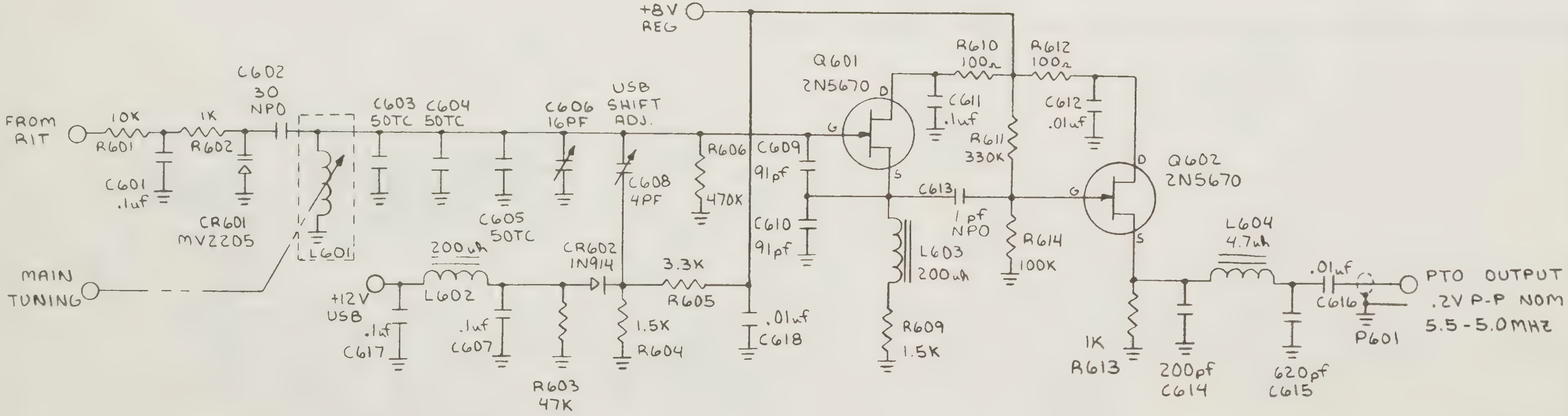
SWAN ELECTRONICS

SCALE: NONE
DATE: 9-14-78
APPROVED BY:
DRAWN BY C DAVIDSON
REVISED

PTO BRD LAYOUT

DRAWING NUMBER
0800-2305

1. COMPONENT DESIGNATORS 7600 SERIES
NOTES



SWAN ELECTRONICS

SCALE: NONE	APPROVED BY:	DRAWN BY C. DAVIDSO
DATE: 9-22-78		REVISED

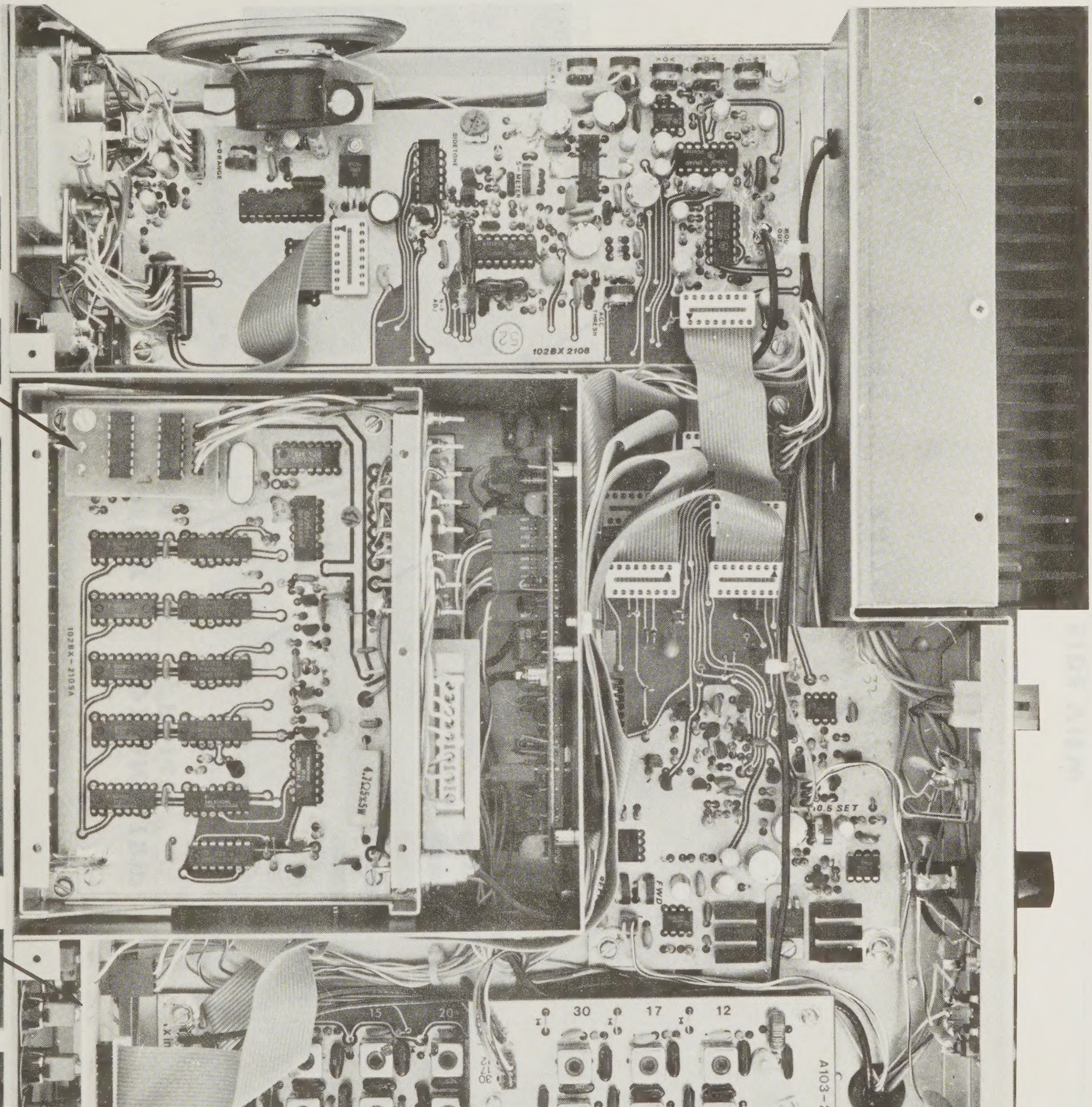
PTO BRD SCHEMATIC (038-188A)

DRAWING NUMBER 0800-2405

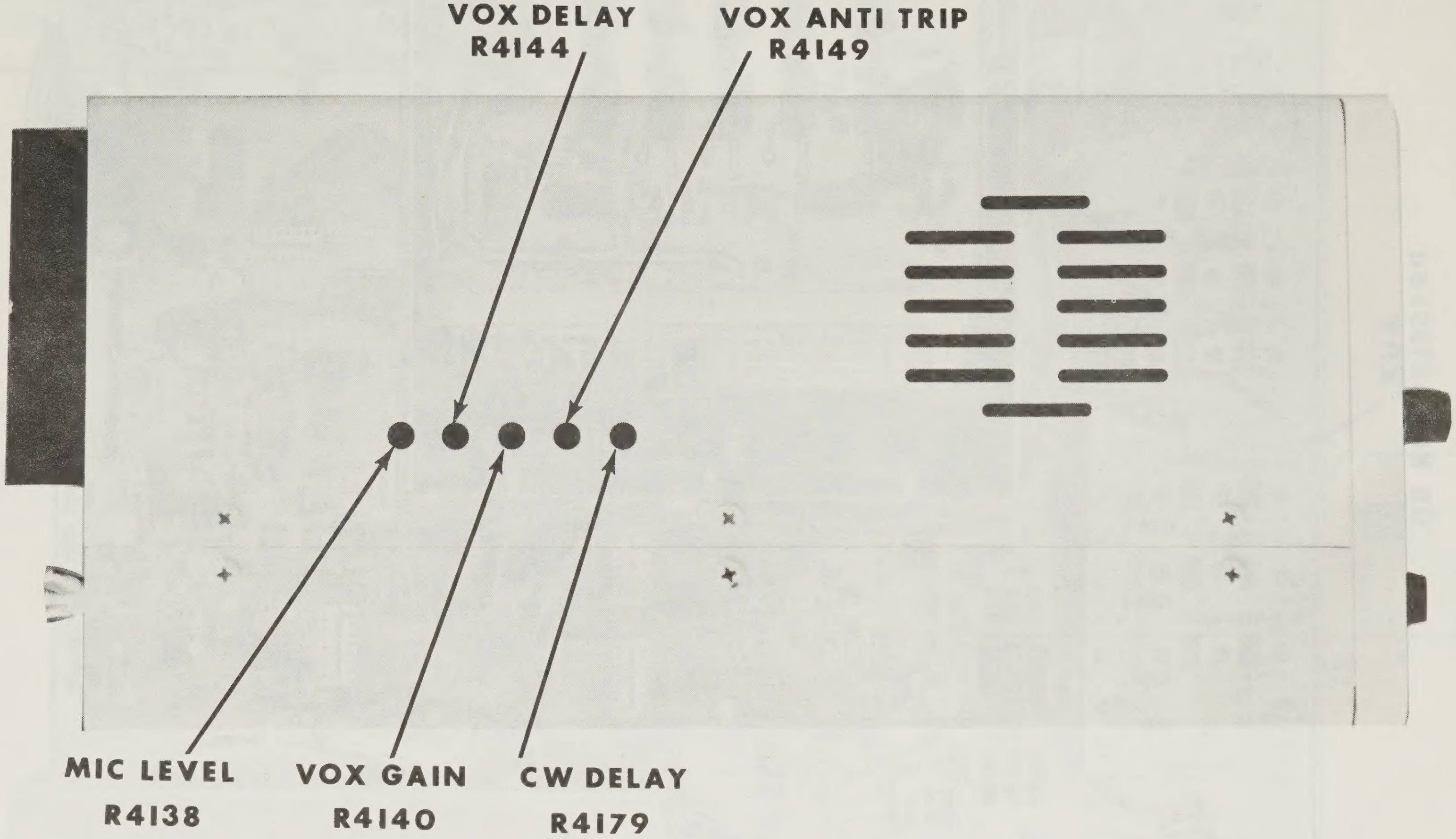
1. COMPONENT DESIGNATORS 7600 SERIES

NOTES

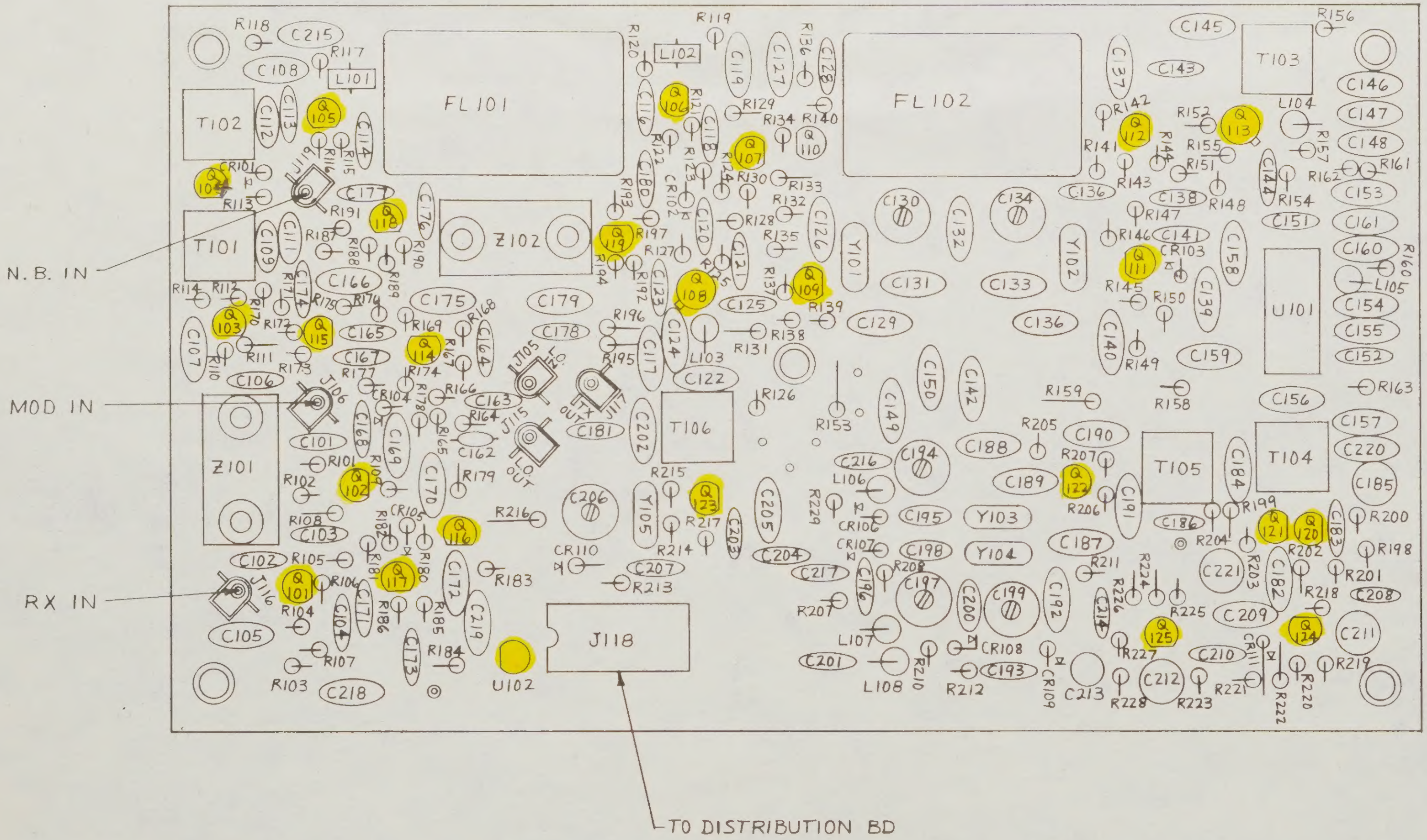
11X17 PRINTED ON NO. 1000H CLEARPRINT


AUX. COUNTER BD.
FRONT PANEL RIGHT BD.
AUX. PRESELECTOR BD.
102BX 2108
102BX-2105A
A103-2102
102BX 2102A
30
17
12
15
20

TOP VIEW



LEFT SIDE VIEW



2. COMPONENT DESIGNATORS 2000 SERIES.
1. REF. ELECT. ASSY 102BX-2007
REF. CIRCUIT BOARD 102BX-2107
REF. DRILL DWG 102BX-2204

NOTES:

SWAN ELECTRONICS

SCALE: NONE
DATE: 9-24-79
APPROVED BY: JW
DRAWN BY AJG
REVISED

EXCITER BOARD LAYOUT

DRAWING NUMBER 102BX-2307